AD NUMBER

AD405676

NEW LIMITATION CHANGE

TO

Approved for public release, distribution unlimited

FROM

Distribution authorized to U.S. Gov't. agencies and their contractors; Administrative/Operational Use; 31 JAN 1963. Other requests shall be referred to Air Force Rome Air Development Center, Griffiss AFB, NY.

AUTHORITY

RADC, USAF ltr, 21 Aug 1964. TAB 64-20 15 Oct 1964

DEFENSE DOCUMENTATION CENTER

FOR

SCIENTIFIC AND TECHNICAL INFORMATION

CAMERON STATION ALEXANDRIA, VIRGINIA

IMPROVED MEASUREMENT TECHNIQUES FOR
SPECTRAL POWER OUTPUT DATA

Technical Note No. 1
31 January 1963

Prepared by

W. K. Brown
J. Dolinski
E. Grier
R. Smith
D. Tang
K. Walker

AEL Report 62046-4

Contract AF30(602)-2677

Prepared for

Rome Air Development Center
Research and Technology Division
Air Force Systems Command
United States Air Force
Griffiss Air Force Base
New York

AMERICAN ELECTRONIC LABORATORIES, INC.
Richardson Road, Colmar, Penna.

PUBLICATION REVIEW

This report has been reviewed and is approved.

Approved: LOUIS F. MOSES
          Task Engineer
          Applied Research Branch

Approved: SAMUEL D. ZACCARI, Chief
          Electromagnetic Vulnerability Lab
          Directorate of Communications

FORWARD

This is an interim report which covers a study and research effort with the goal of increasing the capability of automated interference measurements in the frequency range of 100 kilocycles (kc) to 40 gigocycles (gc).

The work has been performed by American Electronic Laboratories, Colmar, Penna. The technical personnel engaged on this project are, the following: W. Brown, J. Dolinski, E. Grier, R. Smith, D. Tang, and K. Walker.

ABSTRACT

The objective of the study program, "Improved Measurement Techniques for Spectral Power Output Data" is to pursue a continued research and development effort with the goal of increasing the capability of automated interference measurements in the frequency range of 100 kilocycles (kc) to 40 gigacycles (Gc).

Investigation is being directed toward the design, on paper, of a single receiver having the combined capabilities of a spectrum signature, field surveillance, field intensity and other types of broadband receivers. During the course of this study program, major emphasis will be placed on investigation of techniques in two specific areas:

1. voltage-tunable, solid-state receivers, and
2. automated techniques for the presentation.

Where necessary, various techniques will be breadboarded for comparison or feasibility purposes.

The program output will be a final report describing the relative merits of the techniques investigated. The design of a receiver utilizing the techniques which best fulfill the program goals will also be presented on paper.

This program is now approximately six months old and 50% complete.

Field tests were run on the existing AN/GRR-9, a voltage-swept Field Surveillance Receiver. The results of these field tests were applied as inputs to the present receiver study program.

Various receiver approaches were studied and compared. Solid-state techniques are presently being investigated to determine their suitability for implementation in the 100 kc to 40 Gc frequency range. Preliminary results have also been obtained from a study of display devices.

The remainder of the program will be devoted to the study of automated measurement techniques and to the demonstration of an all solid-state feasibility breadboard operating from 2 Gc to 4 Gc. Solid-state receiver front-ends will also be breadboarded between 8 Mc and 16 Mc to demonstrate feasibility of techniques applicable between 100 kc and 250 Mc, were lumped-constant circuits are realizable.

A Technical Note is being submitted at this time since the receiver approach has been finalized.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

LIST OF TABLES

TECHNICAL NOTE

CONTRACT AF30(602)-2677

IMPROVED MEASUREMENT TECHNIQUES FOR

SPECTRAL POWER OUTPUT DATA

1. INTRODUCTION

The objective of the study program, "Improved Measurement Techniques for Spectral Power Output Data," is to pursue a continued research and development effort with the goal of increasing the capability of automated interference measurements in the frequency range of 100 kilocycles (kc) to 40 gigacycles (Gc).

Investigation is being directed toward the conceptual design of a single receiver having the combined capabilities of spectrum signature, field surveillance, field intensity and other types of broadband receivers. A significant portion of the research program is concerned with automated presentation, recording and storage of the data derived from the receiver.

This Technical Note describes the preliminary results of the solid-state receiver investigation.

2. CONCLUSIONS

The voltage-tunable solid state receiver described in this Technical Note is capable of providing Spectrum Surveillance, Spectrum Signature and Field Intensity data.

High receiver sensitivity, variable sweep speed and sweep width, and selectable IF bandwidths provide acquisition of data in the least amount of time for spectrum surveillance. Simultaneous coverage from 100 kc - 40 Gc with variable IF bandwidths provides the capability for obtaining spectrum signatures quickly. Automated techniques for the measurement of amplitude and frequency, to be investigated during the remainder of the program, will allow the receiver to act as an excellent field intensity meter.

Solid state techniques, while causing some degradation of performance over a tube-type receiver, yield a small-size-low-cost receiver which is highly reliable and easily maintainable.

3. RECOMMENDATIONS

No recommendations are applicable to this Technical Note. Recommendations will be made in the Technical Report at the conclusion of the program.

4. TECHNICAL DISCUSSION

A. RECEIVER PERFORMANCE CAPABILITIES[1]

The initial step in the program was to outline performance capabilities of a receiver which will fulfill the program goal of increasing the capability of automated interference measurements in the frequency range of 100 kc to 40 Gc. Moreover, the receiver techniques studies should ideally have the combined capabilities of spectrum signature, field surveillance, field intensity, and other types of broadband receivers.

1. <u>Electrical</u>

    a. Frequency Coverage: 100 kc to 40 Gc

    b. Sensitivity: -90 dbm nominal[2]

    c. Bandwidth: Selectable, depending upon anticipated received signal characteristics.

    d. Sweep Rate[3]: Variable, stop through 1000 cps.

    e. Sweep Width[3]: Variable, to allow for sector scan about any frequency.

    f. Dynamic Range: 60 db minimum

    g. Spurious Signal Rejection[4]: 60 db minimum

    h. Display: Simultaneous for all bands.

    i. Data Storage: (Frequency, Amplitude and Time)

       (1) Preliminary

       (2) Permanent

    j. Antennas:

       (1) Gain accuracy within ± 1 db

---

1. The receiver performance capabilities listed are ideal goals. When actual receiver hardware is considered, tradeoffs and compromises may be necessary in order to produce the optimum practical receiving system.

2. The -90 dbm sensitivity figure is nominal and relates to bandwidths between 1 and 10 Mc. Below 250 Mc, where the information bandwidth is less than 1 Mc, receiver sensitivities will be greater than -90 dbm. Above 10 Gc, where information bandwidths are on the order of 30-50 Mc, receiver sensitivities may be somewhat less than -90 dbm.

3. If a time-shared receiver is utilized. Moreover, the study of detection probability may change this sweep range slightly.

4. Spurious responses include any false signal indications (including the image frequency response) plus any superposition of real or true signals.

4

(2) Broadband
(3) Variable polarization
(4) Omnidirectional and directional

k. Field Intensity Measurements:

(1) Frequency calibration accuracy within 1 part in 10^6 [5]
(2) Amplitude calibration accuracy within ± 3 db overall [5]

2. Mechanical

a. Size and Weight: Capable of mobile van mounting
b. Environmental: Capable of mobile van installation, e.g. MIL-E-4158
c. Mounting: Standard 19-inch rack.

B. RECEIVER COMPARISON

1. General

The purpose of the receiver comparison study was to compare various basic receiver types in order to determine which type or types best fill the requirements imposed by the receiver performance capabilities described previously. Such receiver enhancing devices as preselector filters, low-noise preamplifiers, variable bandwidth I.F.'s, AGC, and AFC, will later be studied on an individual basis.

All signal intercept receivers can be classed into one of two groups: wide-open, or time shared. The wide-open receiver (e.g. the crystal video receiver) will intercept signals anywhere in the receiver passband at any instant of time. The time shared receiver (e.g. manually tuned and panoramic receivers) covers the receiver passband in sequential sector steps.

Wide-open receivers take two basic forms: those providing no frequency resolution aside from the total R.F. bandpass, and those providing features which permit frequency measurement within the receiver passband. The time shared receiver has inherent frequency resolution in the relatively small band increment monitored at any instant of time.

2. Time-Sharing Receivers [6]

Time-sharing receivers employ a narrow band filter which is tuned across the receiver passband in sequential steps. Time sharing devices include the TRF receiver (where a voltage tunable filter is utilized), single and double superheterodyne receivers, and the cyclotron resonance receiver (an electronically tunable device with inherent AM detection).

Tuning may be by manual, mechanical or electrical means.

---

5. The ± 3 db amplitude calibration accuracy can be reduced to ± 1.5 db if automated amplitude readout is not desired.

6. Refer to Appendix I for a more complete description of the various receiving techniques mentioned here.

a. Manual Tuning

Manual tuning is the easiest tuning method to implement. Only slow tuning speeds are possible, but very accurate frequency settings can be achieved.

b. Mechanical Tuning

Manually tuned receivers can easily be adapted to motor-driven tuning. Increased sweep speed is thereby obtained. The manual tuning "mode" is usually retained on mechanically tuned receivers to provide the accurate frequency setting capability.

c. Electronic Tuning

Electronic tuning provides fast receiver tuning speeds. Electronic receiver tuning can be accomplished in two ways: a voltage tunable filter can be tuned in front of a wide-open receiver; or, a voltage tunable local oscillator can be utilized in a superheterodyne receiver. Both techniques have been employed in a single receiver in order to provide increased image and spurious signal rejection.

3. Non-Time Sharing Receivers[7]

Non-time sharing receivers are continuously receptive over the entire receiver frequency range. Two types of receivers have non-time sharing characteristics: crystal video; and, sideband mixing super-heterodyne.

The crystal video receiver is a form of TRF receiver by which any signal within the R.F. bandpass may be received. The sideband mixing superheterodyne is essentially a superheterodyne receiver employing a number of local oscillators spaced in frequency by the I.F. amplifier center frequency. The side-band technique allows 100% intercept probability while providing sensitivities comparable to those of the superheterodyne type.

To provide frequency resolution with a non-time sharing receiver, either channelized filters or frequency discrimination may be utilized.

Channelized, overlapping filters can be placed ahead of wide-open receivers. The individual filter bandwidth determines the frequency resolution.

A frequency discrimination circuit, composed of two R.F. channels, one frequency insensitive and the other having a frequency sensitive attenuator, can also be utilized to provide a means of frequency resolution.

The time-shared and non-time shared receivers described above are compared in Appendix I. A summary of the results is presented in Table I.

---

7. Refer to Appendix I for a more complete description of the various receiving techniques mentioned here.

6

It appears obvious from Table I that the basic receiver system offering the optimum compromise and the most versatility is the single superheterodyne. Spurious and antenna radiation can be suppressed to an adequate level via such techniques as tunable preselector filters, image rejection mixers, and high I.F. frequency. Single intercept probability, sensitivity and selectivity can all be maximized through utilization of a variable sweep rate and variable I.F. bandwidth. Moreover, cost, size, complexity and power comsumption can be minimized by the use of solid state local oscillators.

Ultimately, a non-time shared receiver may be utilized in conjunction with the single superheterodyne receiver in order to provide blanking information for spurious signal rejection. However, such results as this will not become evident until the spurious signal investigation is completed.

C. BAND SELECTION

After determination of a basic receiver type, individual frequency bands must be chosen for receiver implementation[8]. From considerations of signal density and component availability, the following band selections have been made:

Band	Frequency Range	Band	Frequency Range
1	100-250 kc	11	125-250 Mc
2	250-500 kc	12	250-500 Mc
3	500-1000 kc	13	500-1000 Mc
4	1-2 Mc	14	1-2 Gc
5	2-4 Mc	15	2-4 Gc
6	4-8 Mc	16	4-8 Gc
7	8-16 Mc	17	8-12.4 Gc
8	16-32 Mc	18	12.4-18 Gc
9	32-64 Mc	19	18-26.5 Gc
10	64-128 Mc	20	26.5-40 Gc

D. SPURIOUS SIGNAL SUPPRESSION

Spurious or unwanted signals may be divided into six categories:

1. Image signals
2. Strong passband signals
3. Stop-band signals
4. Intermodulation and cross modulation
5. Local-oscillator radiation
6. Intermediate frequency within the local-oscillator bands or signal frequency bands.

A summary of the effects of each spurious response type and the suppression techniques for eliminating these effects are given in Table II.

---

8. Refer to Appendix II for a discussion of receiver band selection.

	SPURIOUS RESPONSES	ANTENNA RADIATION	INTERCEPT PROBABILITY[9]	SENSITIVITY	SELECTIVITY	TYPE
Tuned Radio Frequency (TRF)	None	None	Variable sweep speed, consistent with the bandwidth	Poor without preamplifier, Bandwidth fixed	Relatively poor, band-width fixed	Tube or solid state
Single Superheterodyne	Minimized by: 1. Tunable preselector 2. Image rejection mixer 3. I.F. frequency	Minimized by tunable preselector	Variable sweep speed. The bandwidth can be changed in accordance with the sweep speed.	High, since the bandwidth is variable. Preamplification can be used.	High, vari-able band-width	Tube or solid state
Double Superheterodyne	Minimized by: 1. Tunable preselec-tor 2. Image rejection mixer 3. I.F. frequency	Minimized by: 1. Tunable pre-selector 2. High first I.F. frequency	Variable sweep speed. The bandwidth can be changed in accordance with sweep speed.	Poor unless low noise-high gain preamplifica-tion is used.	High, vari-able band-width	Tube or solid state
Cyclotron Resonance	None	None	Variable sweep, consis-tent with bandwidth and magnetic field change capability.	Poor without preamplifier	Relatively poor, band-width fixed	Tube
Crystal Video	None	None	100%	Poor without preamplifier	None	Tube or solid state
Sideband-Mixing Superheterodyne	Relatively many, due to all the fre-quencies present on the crystal	Relatively high level	100%	Poor without preamplifier. Holes in band.	Relatively poor, band-width fixed	Tube or solid state
Discriminator	None	None	100%	Poor without preamplifier	Relatively poor	Tube or solid state
Fully-Channelized	Ambiguities at crossover frequen-cies	None	100%	Poor without preamplifier	Relatively poor	Tube or solid state

9. An intercept probability of 100% is not necessarily optimum for the spectrum surveillance application. For example, while a sweep rate faster than one microsecond would allow 100% detection probability of a pulse with greater than one microsecond duration, it would be necessary to stop the sweep to see any pulse shape at all. Moreover, receiver bandwidth is now equal to impulse bandwidth, and sensitivity will decrease correspondingly. Therefore, a slow sweep speed, with slightly reduced transient intercept probability, may prove superior due to increased effective bandwidth and sensitivity.

Table I: Receiver Comparison

TIME SHARED RECEIVERS   NON-TIME SHARED RECEIVERS

Table I: Receiver Comparison

	SPURIOUS RESPONSES	ANTENNA RADIATION	INTERCEPT PROBABILITY[9]	SENSITIVITY	SELECTIVITY	TYPE
TIME SHARED RECEIVERS						
Tuned Radio Frequency (TRF)	None	None	Variable sweep speed, consistent with the bandwidth	Poor without preamplifier, Bandwidth fixed	Relatively poor, bandwidth fixed	Tube or solid state
Single Superheterodyne	Minimized by: 1. Tunable preselector 2. Image rejection mixer 3. I.F. frequency	Minimized by tunable preselector	Variable sweep speed. The bandwidth can be changed in accordance with the sweep speed.	High, since the bandwidth is variable. Preamplification can be used.	High, variable bandwidth	Tube or solid state
Double Superheterodyne	Minimized by: 1. Tunable preselector 2. Image rejection mixer 3. I.F. frequency	Minimized by: 1. Tunable preselector 2. High first I.F. frequency	Variable sweep speed. The bandwidth can be changed in accordance with the sweep speed.	Poor unless low noise-high gain preamplification is used.	High, variable bandwidth	Tube or solid state
Cyclotron Resonance	None	None	Variable sweep, consistent with bandwidth and magnetic field change capability.	Poor without preamplifier	Relatively poor, bandwidth fixed	Tube
NON-TIME SHARED RECEIVERS						
Crystal Video	None	None	100%	Poor without preamplifier	None	Tube or solid state
Sideband-Mixing Superheterodyne	Relatively many, due to all the frequencies present on the crystal	Relatively high level	100%	Poor without preamplifier. Holes in band.	Relatively poor, bandwidth fixed	Tube or solid state
Discriminator	None	None	100%	Poor without preamplifier	Relatively poor	Tube or solid state
Fully-Channelized	Ambiguities at crossover frequencies	None	100%	Poor without Preamplifier	Relatively poor	Tube or solid state

9. An intercept probability of 100% is not necessarily optimum for the spectrum surveillance application. For example, while a sweep rate faster than one microsecond would allow 100% detection probability of a pulse with greater than one microsecond duration, it would be necessary to stop the sweep to see any pulse shape at all. Moreover, receiver bandwidth is now equal to impulse bandwidth, and sensitivity will decrease correspondingly. Therefore, a slow sweep speed, with slightly reduced transient intercept probability, may prove superior due to increased effective bandwidth and sensitivity.

As can be seen from Table II, the majority of spurious suppression techniques involve components or circuits outside the fundamental RF "front-end" (mixer, local oscillator and IF amplifier). However, before these techniques can be considered, the intermediate frequency must be properly chosen to eliminate signal or local-oscillator frequency feed-through, and to prevent the generation of harmonic products which produce the intermediate frequency, i.e., $mf_s \pm nf_{LO} = f_{IF}$. Appendix III gives an analysis of the $mf_s \pm nf_{LO}$ products, relative amplitudes, and conditions for generation. From this detailed analysis, local-oscillator frequencies and intermediate frequencies have been determined which allow no spurious responses to be generated for input signal levels below -10 dbm. That is, if a -90 dbm receiver sensitivity existed, no spurious responses would be generated at a level above -90 dbm for all input signal levels up to -10 dbm. A spurious free range of 80 db is therefore achieved.

The acceptable local-oscillator and intermediate frequencies for the bands chosen are given in Table III. Typical receiver RF front-ends utilizing the frequencies specified in Table III are presented in Figure 1.

E. CHOICE OF INTERMEDIATE FREQUENCY

In addition to the spurious response considerations, the intermediate frequency is dependent upon several other factors:

1. Noise figure
2. Criticalness of the circuit
3. Separation of the video from the IF
4. Information bandwidth requirements
5. Noise and spurious responses due to the local oscillator

Each of these factors is discussed fully in Appendix IV. Noise figure and criticalness of circuit are decreased via a low intermediate frequency, and all other factors, including image response, improve with a high intermediate frequency. The choice of intermediate frequency must therefore be a comprise of all factors effected by the intermediate frequency.

F. CHOICE OF RECEIVER FRONT-END

As shown in Table III, three possibilities exist for the local oscillator and IF frequencies in order that 80 db spurious response free operation is achieved. In Case I, the local oscillator frequency range is above the signal frequency range. Case II allows the local oscillator to extend down into the signal frequency range. Case III is essentially a "sub-case" of Case II, except that double conversion is utilized to avoid an excessively high first IF amplifier.

As can readily be seen from Figure 1, implementation of twenty bands utilizing Case I, Case II, or Case III would become quite complex since each band requires different intermediate frequencies. A different IF amplifier, AFC circuit, power-measuring circuit, frequency-

SPURIOUS RESPONSE	EFFECT	SUPPRESSION TECHNIQUE
Image	Receiver responds to two different frequencies for each local-oscillator frequency	1. Preselection 2. High intermediate frequency reduces preselector requirements. 3. Outphase 4. Blanking
Strong passband signal	Receiver saturation	1. Preselector with limiting characteristics 2. Limiter 3. Logarithmic type amplifier 4. Gating
Stop-band signals	Harmonic mixing producing a signal at the intermediate frequency; cross modulation and intermodulation.	1. Preselector filter 2. RF passband filter 3. Proper choice of intermediate frequency 4. Balanced mixer 5. Proper choice of local-oscillator drive power
Intermodulation and cross modulation	Two separate frequencies beat together to produce the intermediate frequency	1. Preselection 2. Minimize circuit non-linearities 3. Gating
Local-oscillator radiation	Spurious response or interference threat to other receivers.	1. Non-reciprocal filter (isolation) 2. RF amplifier with isolation 3. Isolator 4. Preselector 5. Reduction of required local-oscillator drive via tunnel diode converter. 6. Balanced mixer
Intermediate frequency within the fundamental or low harmonics of the signal or local-oscillator frequency bands.	Receiver jamming	1. Proper choice of intermediate frequency

TABLE II.   SPURIOUS RESPONSES AND SUPPRESSION TECHNIQUES

BAND 1: 100-250KC

PRESELECTOR-PREAMPLIFIER 5KC → MIXER → 2.5MC IF 37cps - 3KC

2.6 - 2.75MC L. O.

PRESELECTOR-PREAMPLIFIER 5KC → MI

118 - L.

BAND 2: 250-500KC

PRESELECTOR-PREAMPLIFIER 8KC → MIXER → 5MC IF 62cps - 5KC

5.25 - 5.5MC L. O.

PRESELECTOR-PREAMPLIFIER 8KC → MI

295 - L. (

BAND 3: 500-1000KC

PRESELECTOR-PREAMPLIFIER 15KC → MIXER → 10MC IF 125cps - 10KC

10.5 - 11MC L. O.

PRESELECTOR-PREAMPLIFIER 15KC → MI

595 - L. (

BAND 4: 1-2MC

PRESELECTOR-PREAMPLIFIER 25KC → MIXER → 20MC IF 250cps - 20KC

21 - 22MC L. O.

PRESELECTOR-PREAMPLIFIER 25KC → MI

1.175- L.

BAND 5: 2-4MC

PRESELECTOR-PREAMPLIFIER 50KC → MIXER → 40MC IF .5 - 40KC

42 - 44MC L. O.

PRESELECTOR-PREAMPLIFIER 50KC → MI

2.35 - L.

BAND 6: 4-8MC

PRESELECTOR-PREAMPLIFIER 100KC → MIXER → 80MC IF 1-80KC

64 - 88MC L. O.

PRESELECTOR-PREAMPLIFIER 100KC → MI

4.7 - L. (

NOTE 1: Double superheterodyne techniques must be utilized in Band 7 and above, since the allowable I.F. frequency becomes excessive for practical logarithmic, solid-state I.F. amplifier.

NOTE 2: Above approximately 500 MC the preselector-preamplifier consists of a tunable preselector followed by a wide-open RF preamplifier.

NOTE 3

Figure 1. Receiver Front End Possibiliti

Diagram 1: PRESELECTOR-PREAMPLIFIER 5KC → MIXER → 18KC IF 37cps - 3 KC; L.O. 118 - 545KC

Diagram 2: PRESELECTOR-PREAMPLIFIER 8KC → MIXER → 45KC IF 62cps - 5KC; L.O. 295 - 545KC

Diagram 3: PRESELECTOR-PREAMPLIFIER 15KC → MIXER → 95KC IF 125cps - 10KC; L.O. 595 - 1095KC

Diagram 4: PRESELECTOR-PREAMPLIFIER 25KC → MIXER → 175KC IF 250cps - 20KC; L.O. 1.175-2.175MC

Diagram 5: PRESELECTOR-PREAMPLIFIER 50KC → MIXER → 350KC IF .5 - 40KC; L.O. 2.35 - 4.35MC

Diagram 6: PRESELECTOR-PREAMPLIFIER 100KC → MIXER → 700KC 1 - 80KC; L.O. 4.7 - 8.7MC

the
nsists
llowed
ier.

NOTE 3: The maximum allowable I.F. frequency
         in Case II continues to increase
         above 95 MC. A double superheter-
         odyne receiver shown in Case III
         can therefore, be utilized to min-
         imize the preselector requirements.

NOTE 4: Swept, solid-state local oscillators
         are presently limited to frequencies
         under 40 GC. Therefore, Case I,
         requiring high local oscillator
         frequencies, cannot be utilized
         above the 2 - 4 GC band.

nt End Possibilities (Sheet 1)

11

NOTE 1: Double superheterodyne techniques must be utilized in Band 7 and above, since the allowable I.F. frequency becomes excessive for practical logarithmic, solid-state I.F. amplifier.

NOTE 2: Above approximately 500 MC the preselector-preamplifier consists of a tunable preselector followed by a wide-open RF preamplifier.

NOTE 3: T
i
a
c
c
i

Figure 1.   Receiver Front-End Possibilitie

| PRESELECTOR-PREAMPLIFIER 200 KC | → | MIXER | → | 1.5MC 2 - 160KC |

| | 9.5-17.5MC L. O. | |

| PRESELECTOR-PREAMPLIFIER 400KC | → | MIXER | → | 3.0 MC 4 - 320 KC |

| | 19 - 35 MC L. O. | |

| PRESELECTOR-PREAMPLIFIER 800KC | → | MIXER | → | 6 MC 8 - 640KC |

| | 38 - 70MC L. O. | |

| PRESELECTOR-PREAMPLIFIER 1.6MC | → | MIXER | → | 12 MC 16KC-1.28 MC |

| | 76 - 140MC L. O. | |

| PRESELECTOR-PREAMPLIFIER 3.5MC | → | MIXER | → | 20MC 30KC-2.5MC |

| | 145 - 270MC L. O. | |

| PRESELECTOR-PREAMPLIFIER 7.0MC | → | MIXER | → | 45MC 60KC - 5.0MC |

| | 295 - 545MC L. O. | |

NOTE 3: The maximum allowable I.F. frequency in Case II continues to increase above 95 MC. A double superheterodyne receiver shown in Case III can therefore, be utilized to minimize the preselector requirements.

NOTE 4: Swept, solid-state local oscillators are presently limited to frequencies under 40 GC. Therefore, Case I, requiring high local oscillator frequencies, cannot be utilized

-End Possibilities (Sheet 2)

/ 13: 500-1000MC
Note 2
See Note 3

BAND 14: 1 - 2GC

BAND 15: 2-4GC

NOTE 1: Double superheterodyne techniques
must be utilized in Band 7 and
above, since the allowable I.F.
frequency becomes excessive for
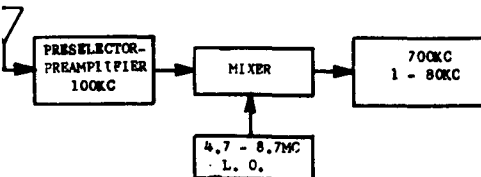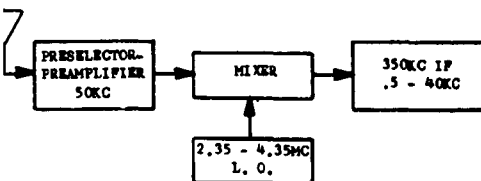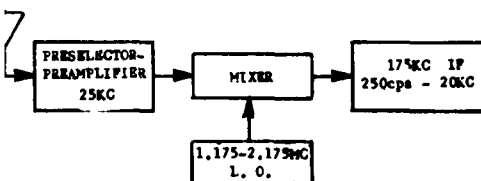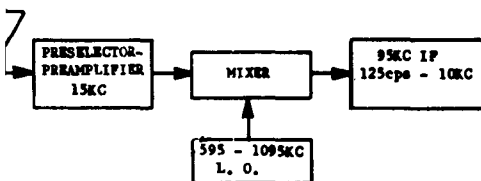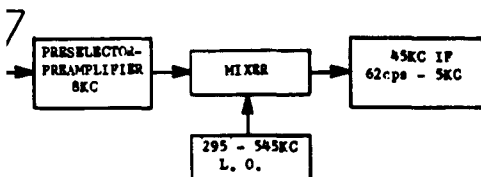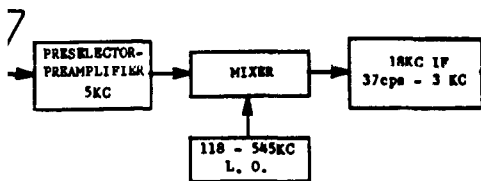practical logarithmic, solid-state
I.F. amplifier.

NOTE 2: Above approximately 500 MC the
preselector-preamplifier consists
of a tunable preselector followed
by a wide-open RF preamplifier.
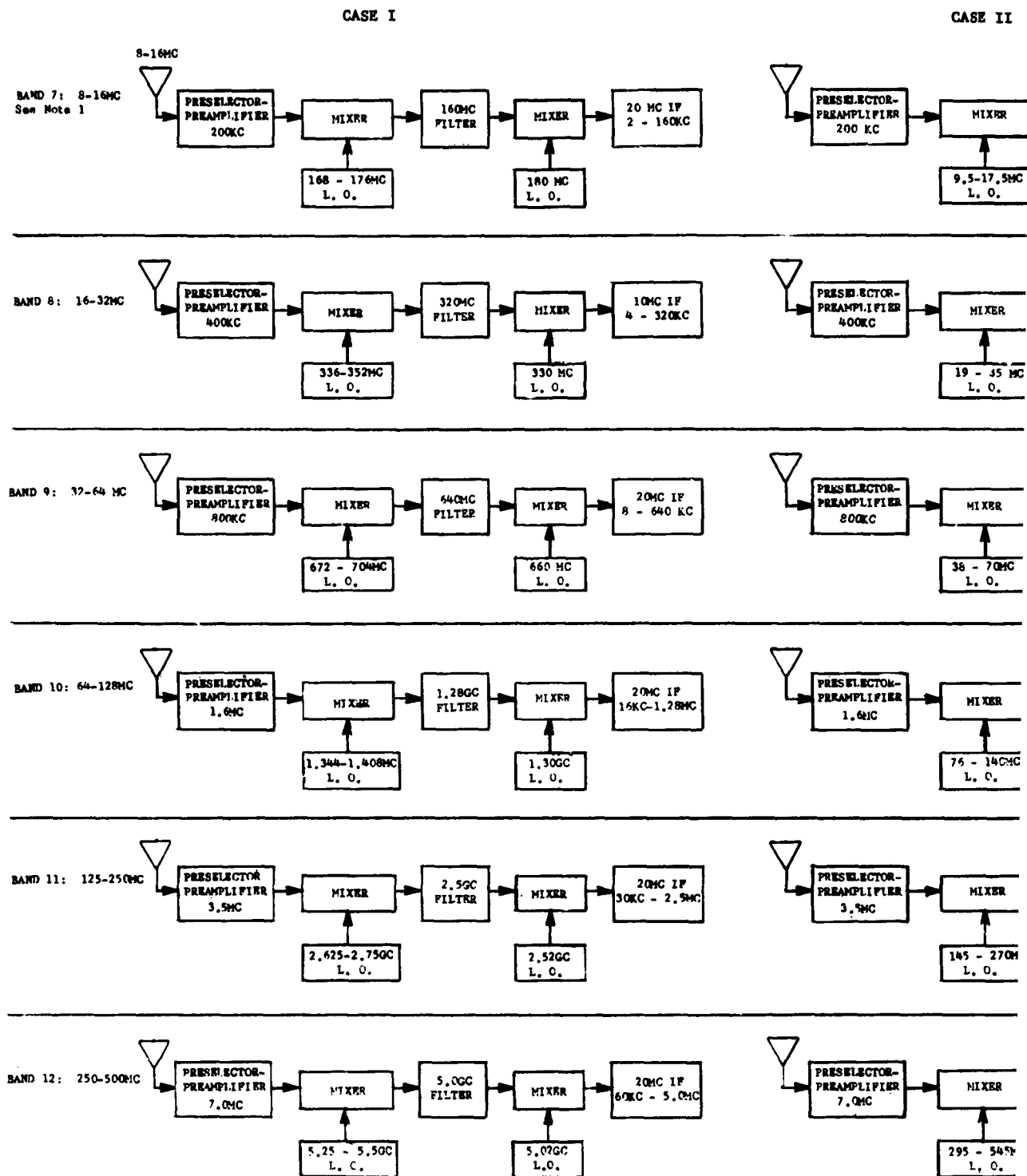
NOTE

NOTE

Figure 1.   Receiver Front-End Possibili

CASE II   CASE III

CASE II — Row 1
PRESELECTOR-PREAMPLIFIER 15.0MC → MIXER → 95MC IF 120KC - 10MC
MIXER L.O.: 595-1095MC L. C.

CASE II — Row 2
PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 95MC IF 240KC- 20MC
MIXER L.O.: 1.1 - 2.1GC L. C.

CASE II — Row 3
PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 95MC IF 240 - 20 MC
MIXER L.O.: 2.1 - 4.1GC L. O.

CASE III — Row 1
PRESELECTOR-PREAMPLIFIER 15.0MC → MIXER → 180MC FILTER → MIXER → 60MC IF 120KC - 10MC
L.O.: 680-1180MC L. C.   L.O.: 240MC L. O.

CASE III — Row 2
PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 380MC FILTER → MIXER → 60MC IF 240KG - 20MC
L.O.: 380 - 2380GC I. O.   L.O.: 440MC L. O.

CASE III — Row 3
PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 750MC FILTER → MIXER → 60MC IF 240KC - 20MC
L.O.: 2.75 - 4.75GC L. O.   L.O.: 810MC L. O.

e techniques   NOTE 3: The maximum allowable I.F. frequency
and 7 and   in Case II continues to increase
wable I.F.   above 95 MC. A double superheter-
essive for   odyne receiver shown in Case III
, solid-state   can therefore, be utilized to min-
                           imize the preselector requirements.

00 MC the   NOTE 4: Swept, solid-state local oscillators
fier consists   are presently limited to frequencies
tor followed   under 40 GC. Therefore, Case I,
amplifier.   requiring high local oscillator
                           frequencies, cannot be utilized
                           above the 2 - 4 GC band.

nt-End Possibilities   (Sheet 3)

BAND 16: 4 - 9GC
See Not 4

PRESELECTOR-
PREAMPLIFIER
25MC

BAND 17: 8-12.4GC

PRESELECTOR-
PREAMPLIFIER
25MC

BAND 18: 12.4-18GC

PRESELECTOR-
PREAMPLIFIER
30MC

BAND 19: 18 - 26.5GC

PRESELECTOR-
PREAMPLIFIER
30MC

BAND 20: 26.5 - 40GC

PRESELECTOR-
PREAMPLIFIER
30MC

NOTE 1: Double superheterodyne techniques
        must be utilized in Band 7 and
        above, since the allowable I.F.
        frequency becomes excessive for
        practical logarithmic, solid-state
        I.F. amplifier

NOTE 2: Above approximately 500 MC the
        preselector-preamplifier consists
        of a tunable preselector followed
        by a wide-open RF preamplifier.
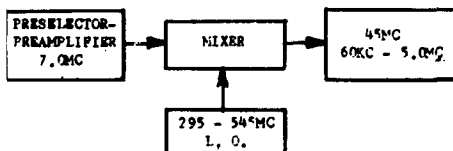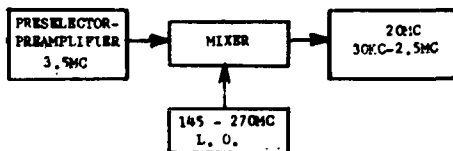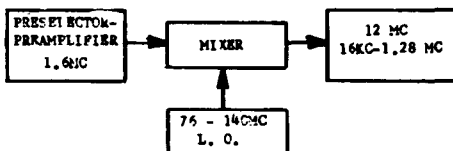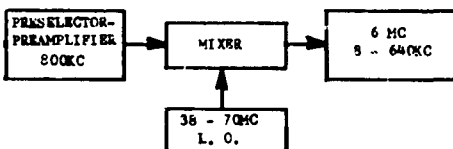
Figure 1.   Receiver Front-End Possib:

CASE II   CASE III

Row 1

CASE II: PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 95MC IF 240KC - 20MC
MIXER L.O.: 4.1 - 8.1GC L.O.

CASE III: PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 1.5GC FILTER → MIXER → 60MC IF 240KC - 20MC
First MIXER L.O.: 5.5 - 9.5GC L.O.; Second MIXER L.O.: 1.56GC L.O.

Row 2

CASE II: PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 95MC IF 240KC - 20MC
MIXER L.O.: 8.1-12.5GC L.O.

CASE III: PRESELECTOR-PREAMPLIFIER 25MC → MIXER → 2.3GC FILTER → MIXER → 60MC IF 240KC - 20MC
First MIXER L.O.: 10.3-14.7GC L.O.; Second MIXER L.O.: 2.36GC L.O.

Row 3

CASE II: PRESELECTOR-PREAMPLIFIER 30MC → MIXER → 95MC IF .5 - 30MC
MIXER L.O.: 12.5 - 18.1GC L.O.

CASE III: PRESELECTOR-PREAMPLIFIER 30MC → MIXER → 3.4GC FILTER → MIXER → 60MC IF .5 - 30MC
First MIXER L.O.: 15.8 - 21.4GC L.O.; Second MIXER L.O.: 3.46GC L.O.

Row 4

CASE II: PRESELECTOR-PREAMPLIFIER 30MC → MIXER → 95MC IF .5 - 30 MC
MIXER L.O.: 18.1 - 26.6GC L.O.

CASE III: PRESELECTOR-PREAMPLIFIER 30MC → MIXER → 5.0GC FILTER → MIXER → 60MC IF .5 - 30MC
First MIXER L.O.: 23 - 31.5GC L.O.; Second MIXER L.O.: 5.06GC L.O.

Row 5

CASE II: PRESELECTOR-PREAMPLIFIER 30MC → MIXER → 95MC IF .5 - 30MC
MIXER L.O.: 26.6 - 40.1GC L.O.

CASE III: PRESELECTOR-PREAMPLIFIER 30MC → MIXER → 7.5GC FILTER → MIXER → 60MC IF .5 - 30MC
First MIXER L.O.: 34 - 47.5GC L.O.; Second MIXER L.O.: 7.56GC L.O.

techniques
nd 7 and
vable I.F.
ssive for
  solid-state
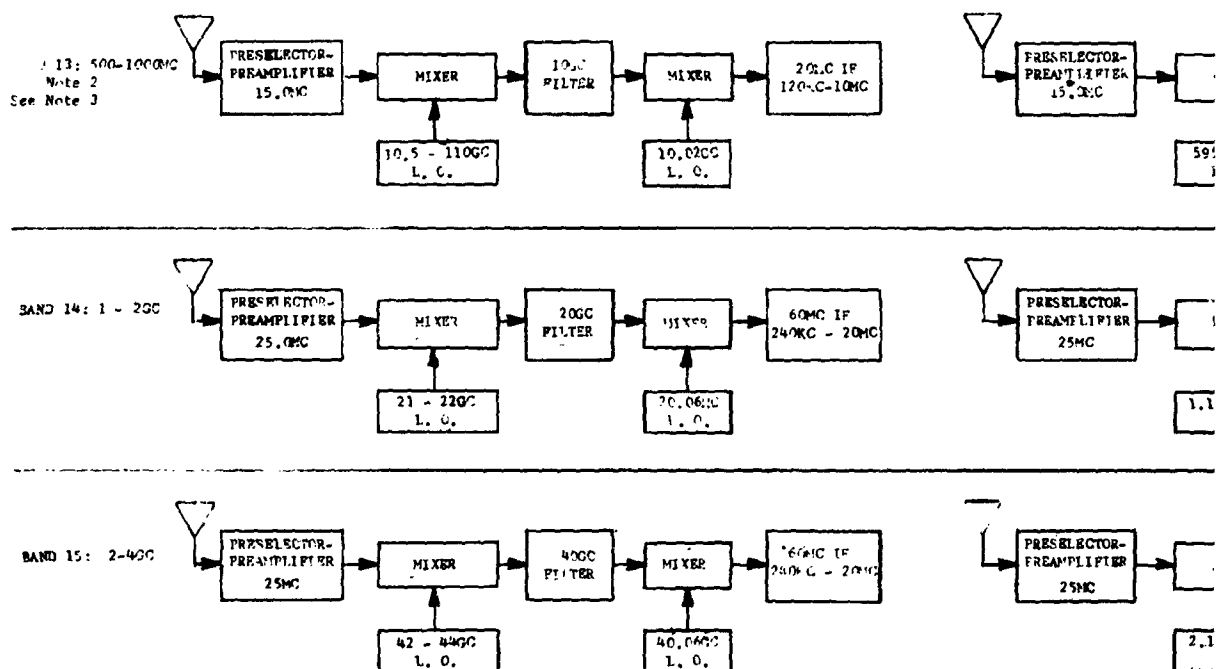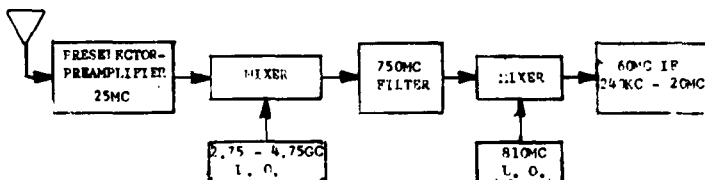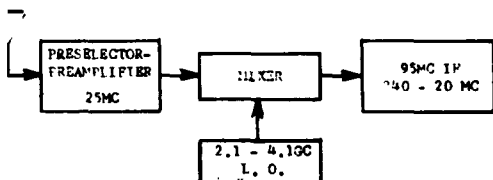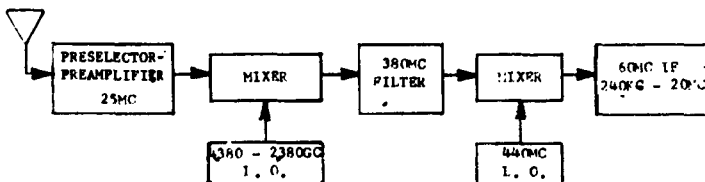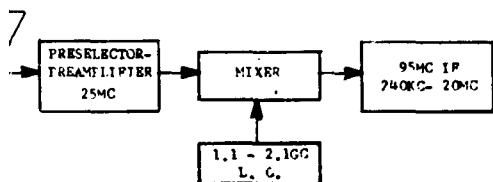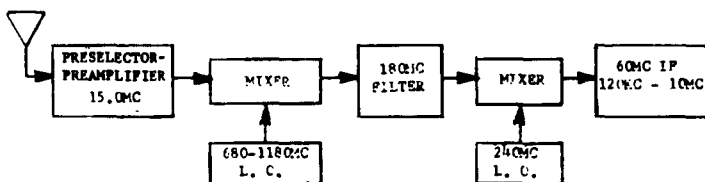
0 MC the
er consists
or followed
mplifier.

NOTE 3: The maximum allowable I.F. frequency in Case II continues to increase above 95 MC. A double superheterodyne receiver shown in Case III can therefore, be utilized to minimize the preselector requirements.

NOTE 4: Swept, solid-state local oscillators are presently limited to frequencies under 40 GC. Therefore, Case I, requiring high local oscillator frequencies, cannot be utilized above the 2 - 4 GC band.

-End Possibilities (Sheet 4)

14

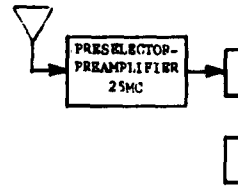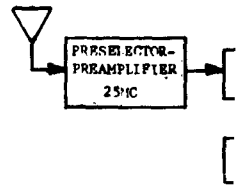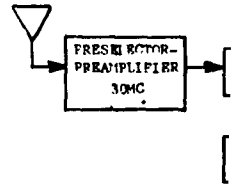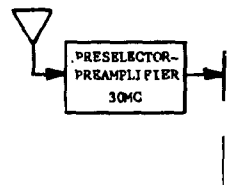| BAND	FREQUENCY RANGE	B + A		B - A			
		MINIMUM L.O. $F_S/F_{LO} < 1/8$	MINIMUM IF $F_S/F_{LO} < 1/8$	MINIMUM L.O. $F_S/F_{LO} < 1/10$	MINIMUM IF $F_S/F_{LO} < 1/10$	MAXIMUM L.O. $F_S/F_{LO} > 5/6$	MAXIMUM IF $F_S/F_{LO} > 5/6$
1	100-250 kc	2.17-2.02 Mc	2.27 Mc	2.40-2.55 Mc	2.50 Mc	119-269 kc	19 kc
2	250-500 kc	4.30-4.05 Mc	4.55 Mc	4.85-5.1 Mc	4.6 Mc	298-548 kc	48 kc
3	500-1000 kc	8.6-8.1 Mc	9.1 Mc	9.6-10.1 Mc	9.1 Mc	596-1096 kc	96 kc
4	1-2 Mc	17.2-16.2 Mc	18.2 Mc	19.2-20.2 Mc	18.2 Mc	1.19-2.19 Mc	190 kc
5	2-4 Mc	34.4-32.4 Mc	36.4 Mc	38.4-40.4 Mc	36.4 Mc	2.38-4.38 Mc	380 kc
6	4-8 Mc	68.8-64.8 Mc	72.8 Mc	76.8-80.8 Mc	72.8 Mc	4.76-8.76 Mc	760 kc
7	8-16 Mc	137.6-129.6 Mc	145.6 Mc	153.6-161.6 Mc	145.6 Mc	9.52-17.52 Mc	1.52 Mc
8	16-32 Mc	275.2-259.2 Mc	291.2 Mc	307.2-323.2 Mc	291.2 Mc	19.04-35.04 Mc	3.04 Mc
9	32-64 Mc	550.4-518.4 Mc	582.4 Mc	614.4-646.4 Mc	582.4 Mc	38.08-70.08 Mc	6.08 Mc
10	64-128 Mc	1.104-1.04 Gc	1.168 Gc	1.236-1.30 Gc	1.172 Gc	76.16-140.16Mc	12.16 Mc
11	125-250 Mc	2.145-2.02 Gc	2.27 Gc	2.425-2.55 Gc	2.30 Gc	149-274 Mc	24 Mc
12	250-500 Mc	4.30-4.05 Gc	4.55 Gc	4.85-5.1 Gc	4.60 Gc	298-548 Mc	48 Mc
13	500-1000 Mc	8.6-8.1 Gc	9.1 Gc	9.6-10.1 Gc	9.1 Gc	596-1096 Mc	96 Mc
14	1-2 Gc	17.2-16.2 Gc	18.2 Gc	19.2-20.2 Gc	18.2 Gc	1.192-2.192Gc	192 Mc
15	2-4 Gc	34.4-32.4 Gc	36.4 Gc	38.4-40.4 Gc	36.4 Gc	2.384-4.384Gc	384 Mc
16	4-8 Gc	68.8-64.8 Gc	72.8 Gc	76.8-80.8 Gc	72.8 Gc	4.768-8.768Gc	768 Mc
17	8-12.4 Gc	137.6-100 Gc	112.4 Gc	120.6-125 Gc	112.6 Gc	9.5-13.9 Gc	1.5 Gc
18	12.4-18 Gc	151.6-146 Gc	164 Gc	176.4-182 Gc	164 Gc	14.8-20.4 Gc	2.4 Gc
19	18-26.5 Gc	223.5-215 Gc	241.5 Gc	259.5-268 Gc	241.5 Gc	21.4-29.9 Gc	3.4 Gc
20	26.5-40 Gc	337.5-324 Gc	364 Gc	390.5-404 Gc	364 Gc	31.5-45 Gc	5 Gc

Table III. Allowable Mixer Inputs for Spurious Products Less than 10th Order.

measuring circuit, and so forth, would therefore be required for each band. Complexity, cost and development time can all be reduced by the utilization of as many common intermediate frequencies as possible. Much time and effort has been spent in the reduction of the number of intermediate frequencies. Considerations involved in minimizing the total number of intermediate frequencies required are described fully in Appendix.V. The frequencies and circuit configurations of Figure 2 represent the optimum compromise of all investigations described previously:

1. Spurious responses
2. Noise figure
3. Criticalness of circuit
4. Separation of video from the IF
5. Information bandwidth requirement
6. Noise and spurious responses due to the local oscillator
7. Minimization of the number of intermediate frequencies

An alternate approach for the receiver implementation is shown in the block diagram of Figure 3. Figure 3 shows a configuration requiring only three different intermediate frequencies. Moreover, only five swept local oscillators are required to cover the complete 100-kc to 40-Gc frequency range, and the highest frequency swept oscillator is 4 to 8 Gc.

The configuration shown in Figure 3 provides only sequential band coverage as opposed to the simultaneous coverage offered by the configuration of Figure 2.

Each circuit represents a different kind of complexity. The configuration represented by Figure 2 requires more equipment (including more swept local oscillators, display devices, automation circuitry duplication, and so forth), thereby becoming complex due to the amount of equipment necessary for total operation. On the other hand, the sequentially swept receiver shown in Figure 3 minimizes the amount of equipment but becomes complex in its operation. In a sequentially swept receiver, complexity is introduced since the operator is required to select the band of operation. This process is more difficult than it initially appears to be. Relying on a priori knowledge only, the operator must first choose which band to use. Next, he must make the more difficult decision of determining the amount of time to allow for the operation of one band while the others are idle. If automatic sequential switching is utilized, valuable information may be lost due to the receiver's remaining on a given frequency band for either too long or too short a period of time.

Aside from the complexity aspects, there are other factors affecting the choice of simultaneous or sequential receiver operation, as shown in Table IV. A review of Table IV shows clearly the superiority of the simultaneous coverage approach shown in Figure 2. The remainder of the program will therefore be devoted to an investigation of techniques for the implementation of the circuit shown in Figure 2.

16

BAND 1: 100 - 250 KC
See Note 1

5 KC PRESELECTOR-PREAMPLIFIER
MIXER
118 - 268 KC L.O.
18 KC I.F. 3 KC
MIXER
63 KC L.O.
45 KC I.F. 37cps - 3 KC

5 KC PRESELECTOR-PREAMPLIFIER
MIXER
10.1 - 10.25MC L.O.
10 MC I.F. 37 cps - 3 KC

BAND 2: 250 - 500 KC
See Note 2

8 KC PRESELECTOR-PREAMPLIFIER
MIXER
295 - 545 KC L.O.
45 KC I.F. 67cps - 5 KC

8 KC PRESELECTOR-PREAMPLIFIER
MIXER
10.25 - 10.5MC L.O.
10 MC I.F. 63cps - 5 KC

BAND 3: 500 - 1000 KC

15 KC PRESELECTOR-PREAMPLIFIER
MIXER
545 - 1045 KC L.O.
45 KC I.F. 125cps - 10 KC

15 KC PRESELECTOR-PREAMPLIFIER
MIXER
10.5 - 11 MC L.O.
10 MC I.F. 125cps - 10 KC

BAND 4: 1 - 2 MC
See Note 3

25 KC PRESELECTOR-PREAMPLIFIER
MIXER
1.175 - 4.175MC L.O.
175 KC I.F. 250cps - 20 KC

See Note 4

BAND 5: 2 - 4 MC

50 KC PRESELECTOR-PREAMPLIFIER
MIXER
2.175 - 4.175MC L.O.
175 KC I.F. .500 - 40 KC

BAND 6: 4 - 8 MC

100 KC PRESELECTOR-PREAMPLIFIER
MIXER
4.7 - 8.7 MC L.O.
700 KC I.F. 80 KC
MIXER
875 KC L.O.
175 KC I.F. 1 - 80 KC

BAND 7: 8 - 16 MC

200 KC PRESELECTOR-PREAMPLIFIER
MIXER
1.5 MC I.F. 2 - 160 KC

BAND 8: 16 - 32 MC

400 KC PRESELECTOR-PREAMPLIFIER → MIXER → 1.5 MC I.F. 4 - 320 KC

17.5 - 33.5 MC L.O.

BAND 9: 32 - 64MC
See Note 3

700 KC PRESELECTOR-PREAMPLIFIER → MIXER → 6.0 MC I.F. 640 KC → MIXER → 1.5 MC I.F. 8 - 640 KC

38 - 70 MC L.O.
7.5 MC L.O.

BAND 10: 64 - 128 MC

1.5 MC PRESELECTOR-PREAMPLIFIER → MIXER → 12.0 MC I.F. 16 KC - 1.28 MC

76 - 140 MC L.O.

BAND 11: 125 - 250 MC

3.5 MC PRESELECTOR-PREAMPLIFIER → MIXER → 12 MC I.F. 30 KC - 2.5 MC

137 - 262 MC L.O.

BAND 12: 250 - 500 MC
See Note 3

7.0 MC PRESELECTOR-PREAMPLIFIER → MIXER → 45 MC I.F. 5 MC → MIXER → 12 MC I.F. 60 KC - 5 MC

295 - 545 MC L.O.
57 MC L.O.

BANDS 13 - 20: Refer to Case II of Figure 2.

NOTE 1: 18 MC is too low an I.F. frequency to be practical. Therefore, double conversion is utilized to produce a 45 MC I.F. frequency.

NOTE 2: A 45 MC I.F. can no longer be utilized due to image rejection requirements.

NOTE 3: A 175 MC I.F. can no longer be utilized due to image rejection requirements. Double conversion to 175 MC is, therefore, utilized.

NOTE 4: Since 175 MC is easily realizable, no necessity exists for continuing the up-conversion to a high I.F. frequency.

Figure 2. Chosen Receiver R-F Front-Ends

17

BAND   FREQUENCY

1   100 KC - 600 KC

2   550 KC - 1050 MC

3   1 MC - 10 MC

4   9 MC - 18 MC

5   17 MC - 26 MC

6   25 MC - 195 MC

BAND PASS FILTER → AMPLIFIER → MIXER (900 KC L.O.)

BAND PASS FILTER → AMPLIFIER → MIXER (450 KC L.O.)

BAND PASS FILTER → AMPLIFIER → MIXER (30 MC L.O.)

BAND PASS FILTER → AMPLIFIER → MIXER (22 MC L.O.)

BAND PASS FILTER → AMPLIFIER → MIXER (16 MC L.O.)

BAND PASS FILTER → AMPLIFIER → MIXER (505 MC L.O.)

1 - 1.5 MC PRESELECTOR-AMPLIFIER → MIXER (3 - 3.5 MC L.O.) → 2 MC I.F. 125 cps-10 KC

31 - 40 MC PRESELECTOR-AMPLIFIER → MIXER (33 - 42 MC L.O.) → 2 MC I.F. 2.5 KC - 200 KC

Figure 3. Alternate Receiver Implementation (Sheet 1)

17 MC - 26 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 16 MC L.O.

25 MC - 195 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 505 MC L.O.

190 MC - 360 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 340 MC L.O.

350 MC - 520 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 480 MC L.O.

510 MC - 680 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 220 MC L.O.

670 MC - 840 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 140 MC L.O.

830 MC - 1000 MC → BAND PASS FILTER → AMPLIFIER → MIXER ← 300 MC L.O.

500 - 1000 MC PRESELECTOR → MIXER ← 560 - 730 MC or 860 - 1030 MC → 30 MC I.F. 50 KC - 4 MC

100 MC I.P.
350 KC - 30 MC

MIXER

4.3 - 5.7 GC
or
6.7 - 8.1 GC

4 - 8 GC
PRESELECTOR

#	Input				L.O.
12	900 MC - 2300 MC	BAND PASS FILTER	AMPLIFIER	MIXER	5.7 GC L.O.
13	2200 MC - 3600 MC	BAND PASS FILTER	AMPLIFIER	MIXER	4.4 GC L.O.
14	3500 MC - 4900 MC	BAND PASS FILTER	AMPLIFIER	MIXER	3.1 GC L.O.
15	4800 MC - 6200 MC	BAND PASS FILTER	AMPLIFIER	MIXER	3.1 GC L.O.
16	6100 MC - 7500 MC	BAND PASS FILTER	AMPLIFIER	MIXER	1.8 GC L.O.
17	7400 MC - 8800 MC	BAND PASS FILTER	AMPLIFIER	MIXER	1.8 GC L.O.
18	8700 MC - 10,100 MC	BAND PASS FILTER	AMPLIFIER	MIXER	3.1 GC L.O.
19	10 GC - 16 GC	BAND PASS FILTER	AMPLIFIER	MIXER	4.4 GC L.O.

Figure 3. Alternate Receiver Implementation (Sheet 2)

ADVANTAGES OF SIMULATANEOUS COVERAGE	ADVANTAGE OF SEQUENTIAL COVERAGE
1. Allows an instantaneous histogram (instantaneous view of all spurious and harmonic outputs). 2. Minimum required time for given surveillance operation. 3. Allows the investigation of more than one signal simultaneously, e.g. intermodulation investigation. 4. Octave band coverage is provided throughout; equal percentage of resolution in each band. 5. High resolution at the low frequencies, where necessary, is available. 6. Allows utilization of standard amplifiers, mixers, wave-guide components, and so forth. 7. Minimizes level of intelligence required by the operator. 8. No a priori knowledge of signal required. 9. No high-frequency preamplifiers are required; therefore, intermodulation is minimized.	1. Minimizes space and cost requirements. 2. Minimizes IF, swept local oscillator and preselector development. 3. Eliminates high-frequency swept local oscillator.

TABLE IV.   SIMULTANEOUS vs. SEQUENTIAL RECEIVER BAND COVERAGE

G. PROBABILITY OF INTERCEPT AND PROBABLE ACQUISITION INTERVAL

1. General Discussion

A "time-shared" receiver is a radio receiver that looks at a given input frequency for a short time and then continues on, successively looking at other frequencies. A "panoramic" receiver is a type of time-shared receiver which automatically and repeatedly tunes across a range of frequencies, presenting its output to an indicator displaying signal amplitude vs. frequency.

Since the panoramic receiver cannot look at all frequencies simultaneously, a problem arises when pulsed RF signals are being investigated. The receiver may not be tuned to the signal frequency at the time the pulse of RF arrives at the receiver. If the emitting antenna rotates, the problem is multiplied. The problem is not peculiar to any one receiver or system.

A concerted effort has been spent in the study of the relationship between the receiver sweep rate, the signal pulse-repetition rate (PRF) and the scan rate of the transmitting antenna for the acquisition of the most data concerning each signal in the least time. A mechanical analog of the probelm is described in Appendix IV. The results described in this report are limited strictly to time and frequency coincidence between signal frequency and receiver tuning. The results presented are not complicated by investigation of the frequency spectrum of pulsed signals or by modifications of receiver bandwidth, which result from the utilization of high sweep rates. The high sweep-rate problem is avoided by limiting the receiver to relatively low scan rates. The assumption of limited bandwidth, which has been made in the following paragraphs, leads to Acquisition Probilities which are less and Maximum Probable Acquisition Intervals which are greater than those which will be actually experienced.

The Intercept Probability reveals the chance of receiving one or more pulses of the signal on a single sweep of the receiver. This information is of minor importance until it is related to the immediate problem. It can be assumed that at least two intercepts (signal acquisitions) are required to be sure a signal actually exists; three or more acquisitions are even better. The time that is really important is the time between intercepts of each signal, since the total of all the intercept times determines the amount of time required to conduct a signal investigation. It has been the goal of this study effort to develop equations for both Unit Coincidence Probability (UPC), which is the probability of seeing a pulse on one random sweep of the receiver, and the Probable Acquisition Interval. These equations are presented in Appendix VI.

It is recognized that pulse-waveform changes result in changes of signal bandwidth, and the acceptance bandwidth of the receiver varies with the sweep rate of the receiver. The equations account for these variables if it is assumed that: (1) a fixed nominal bandwidth of frequencies is contained in the pulse, and (2) that the range of sweep rates is within a region where any variation due to sweep rate is infinitesimal.

21

2. <u>Summary of Conclusions</u>

The study and investigation of Probability of Intercept and Probable Acquisition Interval has yielded numerous results. A summary of the conclusions obtained from the study follows:

a. The problem of intercept involves only pulses or intermittent signals. Continuous-wave signals above noise are received on every sweep.

b. For initial acquisition of signals, the overcoming of synchronous effects requires the use of time-jittered sweep-starting, or time-jittered sweep duration. The former would probably be the easier to implement.

c. The one receiver parameter that has the most effect on acquisition interval is the ratio

$$\frac{kb_r}{B_r} = \frac{\text{Active sweep factor x instantaneous receiver bandwidth}}{\text{Frequency range swept by the receiver}}$$

High resolution and high signal-to-noise ratios require small values of B_r, while efficient use of equipment suggests large values of B_r. The result may be very long acquisition times.

d. A slow receiver-frequency sweep-speed sets a low limit on the time interval between signal acquisitions. Any number of pulses achieved on each sweep can be seen only once per sweep, at intervals of $1/f_r$ second (where f_r is the receiver scan frequency).

e. To achieve the shortest possible interval between acquisitions of pulse signals from rotating antennas, the shortest possible acquisition interval for pulse signals typical of the RF range under observation must be achieved.

f. When there is doubt of actual presence of a signal which occasionally produces an indication, going immediately to a narrow sector-scan will shorten the time required to verify the signal presence. Indeed, the most rapid study will be obtained if the strong signals are initially ignored and all effort is concentrated on the spaces between strong signals.

g. The raw probability equations can be used only to tell the chance of seeing a known signal on one random sweep of the receiver. More important to this investigation is to find how often each signal can be expected to be acquired.

h. As a result of this study the sweep speeds considered optimum are the range from the threshold of noticeable flicker, about 15 cps, up to the slowest PRF expected for the receiver band in use. The lowest expected PRF for ranges up through X-band may be 50 to 200 cycles, so sweep speeds of 15 to 60 cps may be optimum. At much higher bands the short effective range and high accuracy required suggest the PRF would be usually above 1000 pps, and the desirable sweep speeds will range from 15 to 1000 cps.

i. A permanent-recording indicator is made necessary by the long acquisition intervals resulting from slowly scanned antennas and the use of receivers having small values of $k(b_r/B_r)$ to get good sensitivity

and good resolution. An operator might otherwise need to stare without blinking for minutes, at the risk of missing an elusive acquisition.

j. The time interval between acquisitions of a given signal will be the same as the critical sweep period. This is an apparent conclusion which requires further investigation and study. An equation makes it possible to find the fastest sweep which will acquire a pulse from a given signal on every sweep. Call that sweep frequency f_{rm}. Then $T_{rm} = 1/f_{rm}$ is the time interval occupied by each sweep, and fixes the interval between acquisitions of the signal.

On the basis of the probability equations alone, if $T'_H = 1/p^{(1)}$, the approximate time interval between successive acquisitions of a given signal will be the same as the critical sweep period, T_{rm}. A report covering further study of this problem will be completed at a later time.

H. UNIT SPECIFICATIONS

A summary of preliminary unit specifications for the RF front-end components is given in Table V. A complete derivation of these specifications is given in Appendix VII. The specifications listed in Table V are ideal, and are therefore subject to minor changes when practical hardware implementation is considered.

I. RF FRONT-END: 100 KC-250 MC

Since both the preselectors and the oscillators in the 100-kc to 250-Mc range are of the lumped constant type, and since both rely on the changing of either of two variables for a frequency change, the frequency-tuning discussion which follows applied both to preselectors and oscillators.

The preselector bandwidth and tunable frequency range, together with the local-oscillator frequency range, are given in Table V.

As can be seen from the expression for frequency, $f = 1/2\pi\sqrt{LC}$, only two quantities, either the capacitance or the inductance, can be changed in order to vary the frequency over an octave range.

Electrically tuned capacitors (varicaps) and inductors (increductors) are available, but both have inherent characteristics which must be compensated for if reliable operation is to be achieved. Both are relatively low Q components; therefore, the use of these components in a circuit requires special consideration in the design of the rest of the circuit if a degree of selectivity is to be achieved. The range of tuning, especially for the electronically tuned capacitor, is somewhat limited; therefore, care must be taken to reduce the circuit capacitance to a minimum. The capacitors are voltage-driven devices and the inductors are current driven. The maximum frequency which can be obtained by the utilization of an electrically tuned capacitor is approximately 200 Mc; the maximum frequency obtainable with the inductor is approximately 400 Mc.

TABLE IV SUMMARY OF UNIT SPECIFICATIONS

	PRESELECTOR		LOCAL OSCILLATOR				IF AMPLIFIER			
Signal Band Frequency	Band-width	Number of Sections	Frequency	Power Out	Sweep Speed	Stability (w/AGC; w/o AGC)	Band-width	Center Frequency	Characteristics	Approx. Gain
100-250 kc	5 kc	4	118-268 kc	Sufficient for Optimum Operation of Mixer.	15-60 cps ; 50-1000 cps	w/AGC: Approx. 1 Part in 10^6. w/o AGC: Approx. 1 Part in 10^4.	37 cps-3 kc	45 kc	Linear for the first 10 db-20 db of Gain. Logarithmic after linear.	Sufficient to detect the Noise Power; approx. 100 db.
250-500 kc	8 kc	4	295-545 kc				67 cps-5 kc	45 kc		
500-1000 kc	15 kc	4	545-1045 kc				125 cps-10 kc	45 kc		
1-2 Mc	25 kc	4	1.175-2.175 Mc				250 cps-20 kc	175 kc		
2-4 Mc	50 kc	4	2.175-4.175 Mc				500 cps-40 kc	175 kc		
4-8 Mc	100 kc	4	4.7-8.7 Mc				1-80 kc	175 kc		
8-16 Mc	200 kc	4	9.5-17.5 Mc				2-160 kc	1.5 Mc		
16-32 Mc	400 kc	4	17.5-33.5 Mc				4-320 kc	1.5 Mc		
32-64 Mc	700 kc	4	38-70 Mc				8-640 kc	1.5 Mc		
64-128 Mc	1.5 Mc	4	76-140 Mc				16 kc-1.28 Mc	12 Mc		
125-250 Mc	3.5 Mc	4	137-262 Mc				30 kc-2.5 Mc	12 Mc		
250-500 Mc	7.0 Mc	4	295-545 Mc				60 kc-5 Mc	12 Mc		
500-1000 Mc	15.0 Mc	4	595-1095 Mc				120 kc-10 Mc	95 Mc		
1-2 Gc	25.0 Mc	4	1.095-2.095 Gc				240 kc-20 Mc	95 Mc		
2-4 Gc	25.0 Mc	4	2.095-4.095 Gc				240 kc-20 Mc	95 Mc		
4-8 Gc	25.0 Mc	4	4.095-8.095 Gc				240 kc-20 Mc	95 Mc		
8-12.4 Gc	25.0 Mc	4	8.095-12.495 Gc				240 kc-20 Mc	95 Mc		
12.4-18 Gc	30.0 Mc	4	12.495-18.095 Gc				.5-30 Mc	95 Mc		
18-26.5 Gc	30.0 Mc	4	18.095-26.595 Gc				.5-30 Mc	95 Mc		
26.5-40 Gc	30.0 Mc	4	26.595-40.095 Gc				.5-30 Mc	95 Mc		

Feasibility breadboards have been built in the 8 Mc-16 Mc frequency band utilizing both electronically tuned capacitors and inductors. Photographs of these breadboards are shown in Figures 4 and 5. Circuit details and preliminary test results are presented in Appendix VIII.

J. RF FRONT-END: 250 MC-40 GC

As shown in paragraph I, techniques which can be utilized for the development of preselectors, preamplifiers and local oscillator-convertors are identical in the 100-kc to 250-Mc region, where lumped-constant circuits are realizable. However, techniques in the 250-Mc to 40-Gc range differ and each circuit block must be discussed separately.

1. RF Preselector

Since the receiver under consideration covers 100 kc-40 Gc, it is mandatory that a high level of preselection be utilized. Two types of preselection are necessary: one, to reject signals outside the total receiver tunable bandwidth; and another, to reject signals within the receiver tunable bandwidth, but outside the instantaneous receiver (IF) passband.

Band-pass filters can be utilized to provide rejection of signals outside the tunable band. However, most band-pass filters exhibit spurious responses at the 2nd or 3rd harmonic. Therefore, it is proposed to use complementary high- and low-pass filters. Extremely sharp selectivity can be obtained (60-db attenuation at ±5 percent away from the passband), and no spurious responses are evident up to the 7th or 8th harmonic.

Ferrite, current-tunable preselectors can be utilized for the rejection of signals which are within the tunable bandwidth, but outside the instantaneous passband. These filters, employing lithium ferrite or yittrium-iron-garnet (YIG) crystals, are available between 160 Mc and 85 Gc. Characteristics of YIG filters which are advantageous for the present preselection application are:

1. Low insertion loss
2. Narrow instantaneous bandwidth
3. Low limiting level
4. Frequency-selective limiting (one limiting signal does not cause desensitization of the entire band).
5. Minimum phase distortion while limiting
6. Multi-section tracking
7. Isolation (non-reciprocity)

Two double-tuned yittrium-iron-garnet (YIG) filters have been ordered from Watkins-Johnson for incorporation into the 2-4 Gc breadboard. These filters will meet the following specifications:

1. Tuning range: 2-4 Gc
2. Bandwidth: 25 Mc
3. Bandwidth variation with frequency: 2:1 max.
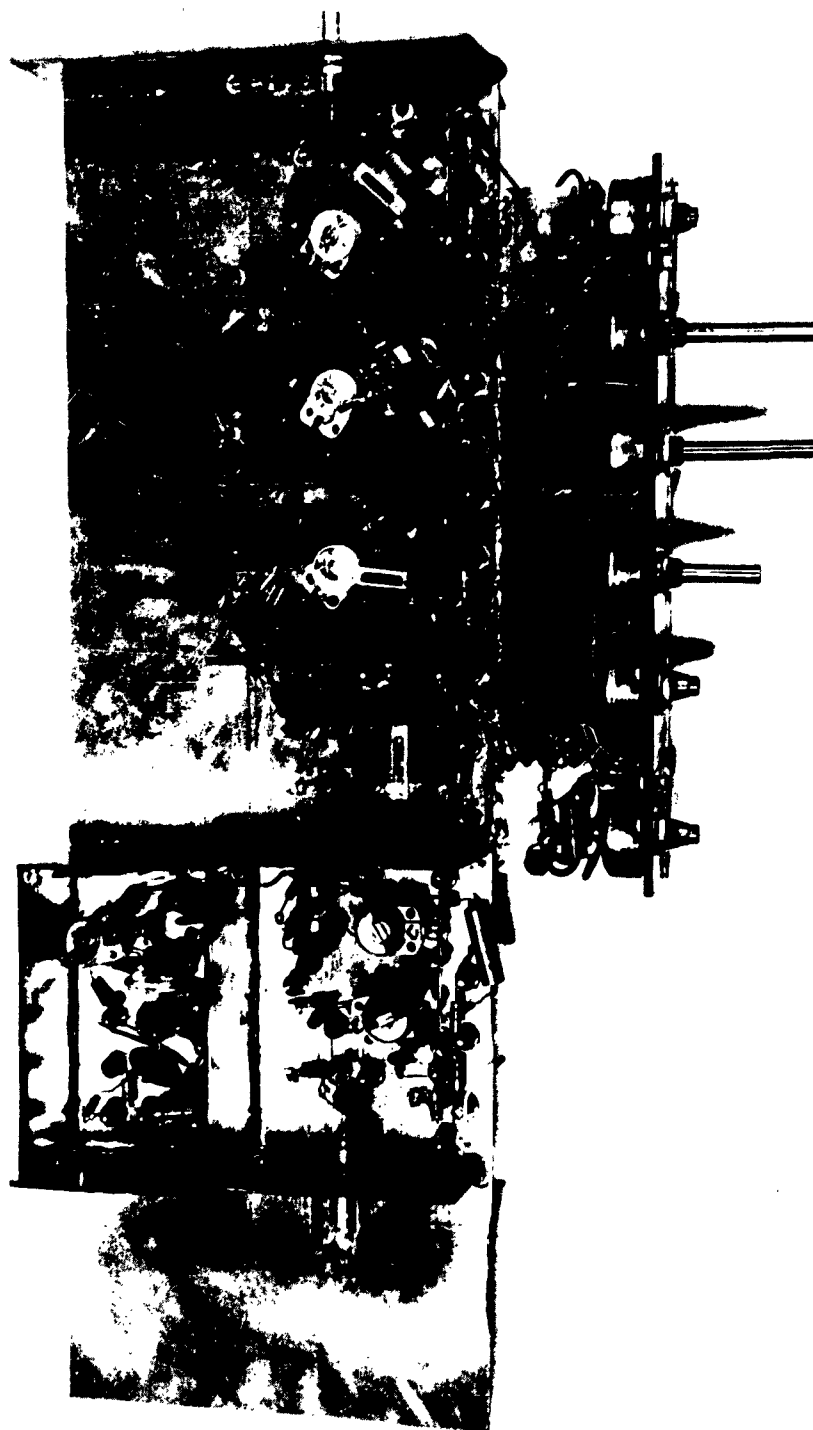4. Insertion loss: 2 db

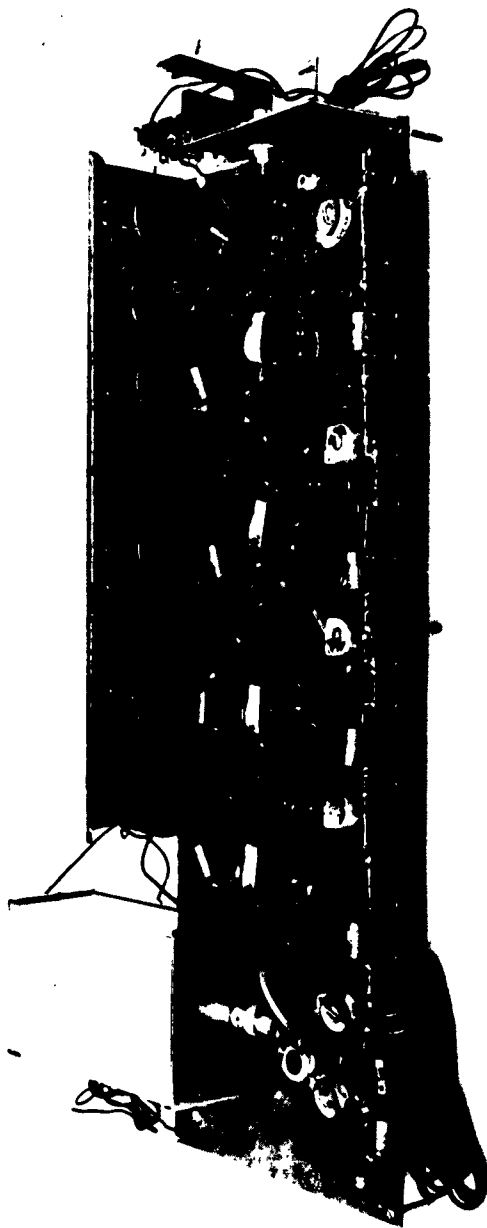25

Figure 4. 8-16 Mc Varicap-Tuned Front-End

4664

Figure 5. 8-16 Mc Increductor Tuned Front End

5. Off resonance isolation: 50 db min.
6. Selectivity: 12 db/octave minimum
7. DC tuning power: 5 watts max.
8. Switching time: 10^{-3} sec. max.
9. Spurious response: 30 db
10. Tracking accuracy: deviation not greater than 5 Mc at any point in the 2-4 Gc band.
11. Hysteresis: deviation not greater than 5 Mc
12. Sweep source: capable of being driven by a transistorized sweep source.

The off resonance isolation of 50 db will be increased to 100 db by the utilization of an RF amplifier between the two filters. Incorporation of an amplifier between the filters will also double the DB rejection of the spurious responses. Work is presently underway toward the design of transistorized sweep and tracking circuitry for the RF preselectors.

A detailed discussion of the preselectors and their relative advantages and disadvantages is presented in Appendix VIII.

2. RF Preamplifier

Tube- and transistor-type amplifiers operating up to a few Gc's are available. These units are within the state-of-the-art, and their properties are well known. No effort will be spent in the investigation of these devices under the current program.

Tunnel diode amplifiers are beginning to present interesting possibilities at microwave frequencies. Units are now commercially available up to 3 or 4 Gc. Certainly no fundamental limitation exists up to 40 Gc.

A theoretical analysis of a tunnel diode amplifier is given in Appendix IX. During the next quarter of the program, a tunnel diode amplifier will be constructed for the 2- to 4-Gc band.

3. Local Oscillator-Converter

Although primary emphasis on the local-oscillator study program has been centered on tunnel diode devices, since solid-state implementation is desirable, some investigation of tube-type oscillators has been done.

a. Backward-Wave Oscillators

Backward-wave oscillators and voltage-tunable magnetrons are available over the entire 250-Mc to 40-Gc range. However, many systems have been built with these devices, and AEL is aware of the advantages and disadvantages of their use. Therefore, only minimal effort has been expended in this area.

b. Backward-Wave Converter

Another tube-type oscillator investigated was the backward-wave converter(BWC). The BWC is a traveling-wave device which combines in one envelope a backward-wave amplifier, a backward-wave oscillator

28

and a mixing structure.

In a program for Patrick Air Force Base (AF08(606)-4660), IT&T Laboratories of Nutley, N.J., is engaged in the refinement and packaging of BWC and BWA tubes originally developed for Rome Air Development Center (AF30(602)-2025). The major emphasis of the current program is wide dynamic range. The aim is for a dynamic range from -80 dbm to +36 dbm.

A detailed discussion of the BWC tube and its characteristics is presented in Appendix IX. A summary of the major BWC properties is given below:

Available Frequencies (Megacycles):	Size: L x D (Inches)
500 - 900	55-1/2" x 3-1/4"
850 - 1500	49-1/2" x 2-1/2"
1450 0 2650	41" x 2"
2500 - 4000	33" x 1-3/8"
3850 - 6000	27-1/2" x 1-1/8"
5700 - 8400	21" x 7/8"
8000 - 12,000	19" x 7/8"

Image Rejection:	35 db minimum
Noise Figure:	20 db
Gain Flatness:	± 3 db over band for a given tube
Dynamic Range:	-80 dbm to +36 dbm goal
Sweep rate:	Up to 100 kc at X-band
	Up to 1 kc at L-Band.
Life:	1000-2000 hours
Environmental Aspects:	Ground-based equipment only
Focus:	Solenoid
Cooling:	Water

No further investigation of these devices will be conducted during this program. Many of the specifications listed above are unsatisfactory for the present application. Moreover, commerical availability of tubes above 12 Gc cannot be expected for at least 2 years.

C. Tunnel Diode Converter

The practicality of tunnel diode oscillators between 250 Mc and 40 Gc has been studied. Esaki or tunnel diode oscillators operating at microwave frequencies up through the X-band have been reported in current literature. Fundamental power up to 40 Gc has been reported by Trambarulo and Burrus of Bell Laboratories.[10] Output powers in the microwatt level (sufficient to drive a tunnel diode converter) and tunability of a few percent are also recorded.

The results of a preliminary tunnel diode oscillator investigation are presented in Appendix VIII. The autodyne or down-converter oscillator will be built and compared to the separate local oscillator and mixer. Oscillator circuits with and without circulators will be tested,

10. R. Trambarulo and C. A. Burrus, "Esaki Diode Oscillators from 3 to 40 KMC," Proceedings of the IRE, Vol. 48, (October, 1960),pp. 1776-1777.

as well as hybrid-coupled mixer circuit.

Tunnel diode oscillators and converters definitely show promise as local oscillators and mixers for the present application. Figure 6 is a photograph of a tunnel diode oscillator tunable from 1 Gc-2 Gc. A second breadboard oscillator which will be electronically tunable from 2 Gc-4 Gc is presently under construction.

K. INDICATING DEVICES FOR SURVEILLANCE RECEIVERS

The type of indicator which is proper for each surveillance receiver depends on the type of program. Typical uses may be field surveillance, countermeasures, spectrum signature, and channel monitoring. Countermeasures receivers require immediate indication for analysis of signal intercept. Field surveillance and spectrum signature studies require both instantaneous readout and data recording. Channel monitoring in many cases requires only recording of data.

As a result of the studies conducted under this program the decision has been made to recommend a memory-type cathode ray tube as the indicating device. Permanent records will be made by photographing the face of the CRT. The following paragraphs detail some of the studies and discuss actual tests which were performed. The tests described in paragraph D-4 proved the utility of the falling raster scan to improve signal identification, and demonstrate techniques for achieving short Acquisition Interval. A block daigram of the proposed indication system is shown in Figure 7.
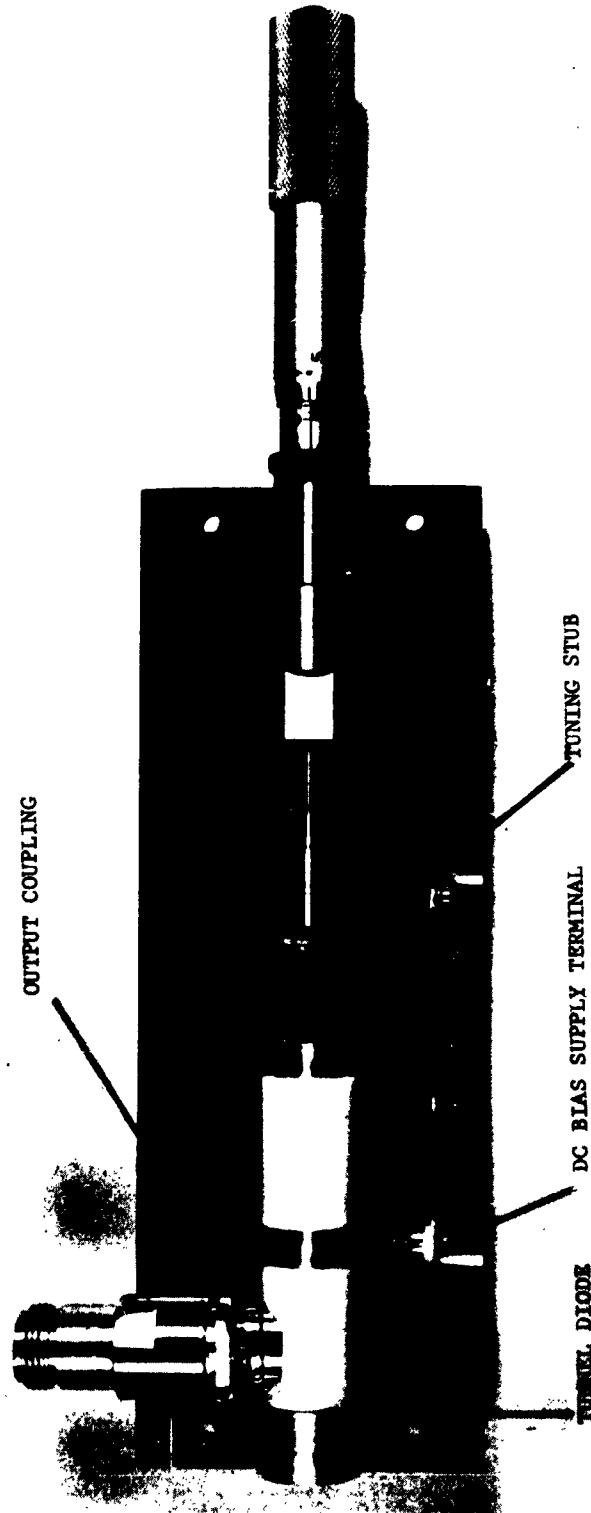
1. Indicating Devices

Several types of indicating devices have been investigated, the principle one of which has been cathode ray tube devices. Some investigation and experimentation has been carried out with chart recorders which would record signal intensity, frequency and time of occurence. Dry chart papers were found, for the most part, to be unsuitable. Moist chart papers have been found somewhat awkward to handle but further investigation is being pursued. Photographic type chart paper for recorders is under investigation but the photographic process of direct recording by CRT display or fiberoptic devices appears to be too slow and cumbersome for the normal RFI program.

a. Cathode Ray Tubes

The effort has therefore been concentrated principally on determining the optimum cathode ray tube device. Ordinary cathode ray tubes with standard phosphors have been tried using P1, P4, P7, and P11 phosphors. All but the P7 phosphor are standard, medium or short persistence type. The P7 phosphor was investigated in an effort to determine whether the phosphorescent storage characteristic of the phosphor would allow the operator to more easily identify signals which appear as short pulses only at infrequent intervals. During an experiment carried on by AEL personnel at RADC, operating in the test van under actual field conditions, it was found that the P7 phosphor produced some improvement in visibility if the van were totally darkened.
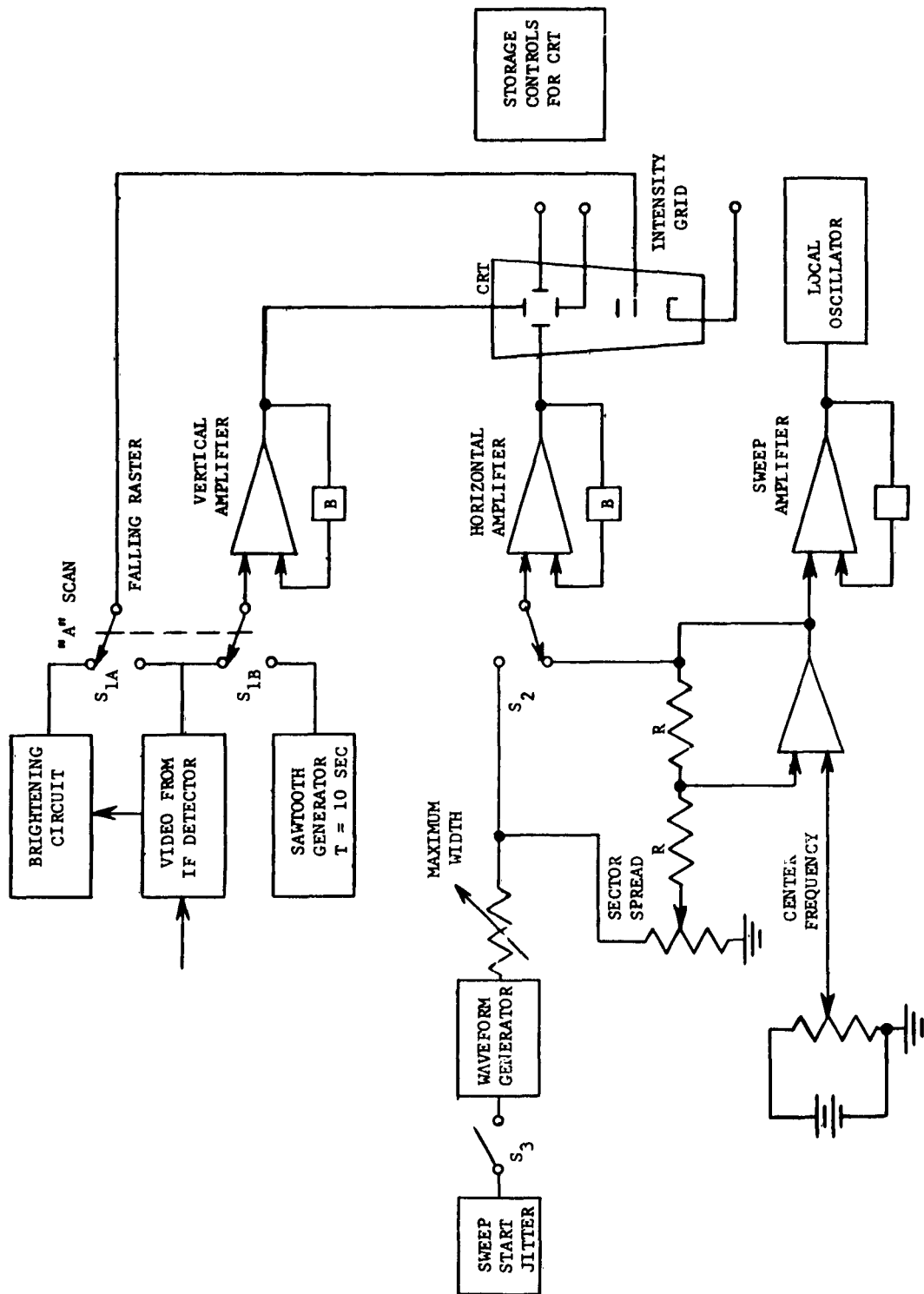
30

461U

OUTPUT COUPLING

TUNNEL DIODE

DC BIAS SUPPLY TERMINAL

TUNING STUB

31

Figure 7. Proposed Indicator System

32

Under room lighting conditions, the long persistent phosphor actually was less visible than the standard short persistent phosphors for short pulse signals.

    b.   <u>Image-Storing CRT</u>

Image storing devices may be thought of as short-term recording devices. The storage-type cathode-ray tube has been investigated, acutal tests performed and photographic records made. This type indicator is recommended as the most important type applicable to field surveillance and spectrum signature.

Memory-type cathode ray tubes, also called storage tubes, were investigated and some experiments were conducted with a Hughes Memoscope Model 106 attached to the indicator of the AN/GRR-9 (XW-1). Very significant improvement of signal acquisition was obtained with the memory device. Even the shortest burst of signal can be recorded and retained until the operator has time to investigate. It must be noted however, that one disadvantage of this type of tube when used for an A-scope display is that a strong noise burst may have the same appearance as a signal.

An illustrated discussion of the study and analysis is presented in paragraph K-4.

For the receiver being developed in the present program, the recommended indicating device is a storage type cathode-ray-tube such as the Hughes "Memotron", ITT's Iatron, or Matchlett Labs., or others. These tubes can write, store, and display rapid pulse information when required, or by the simple resetting of a control they can be made non-storing. The presentation will be switchable from type A scan to a type C (falling raster) scan. It is recommended that each channel of the panoramic receiver may have its own small standard memory type tube with instantaneous writing option and that a large auxiliary Tonotron (the Hughes brand name) type of tube, which allows instantaneous writing and storage simultaneously, may be used as a master analyzing instrument capable of being switched or patched to any channel of the receiver.

There is some background storage which is a characteristic of memory tubes and from time to time the screen must be washed clean of the ghosts of previous traces. For the day-to-day investigation this is not a serious problem, but the characteristic does prevent the memory tube from being a truly long term storage device where information may be desired to be integrated over several hours or more.

    2.   <u>Combined Indicator-Recorder Devices</u>

The paper chart recorder exhibits the advantage of giving both an instantaneous indication and a permanent record of acquired data. There are several photographic processes that allow development and readout within one minute. Nearly any cathode-ray-tube indicator may be photographed intermittently or continuously through a dichroic mirror. In addition, there are devices using magnetic inks and

others with special photographic paper which are used in computer and data processing applications.

a. Paper Chart Recording

The normal sweep rate and acquired signal duration render the normal moving stylus chart recorder useless for spectrum recording. There is the added problem that the receiver sweeps all appear end-to-end, resulting in long strips to be analyzed for signal data.

By altering the mode of operation, the utility of a chart recorder can be improved. The method recommended is very similar to that used for facsimile recorders. The stylus may be scanned across the paper in synchronism with the receiver frequency sweep while the paper advances slowly. Then some means of electrically marking the paper is utilized. Each signal will produce a dot on the paper. This type of recording has the advantage of aiding the separation of signal information from random noise by allowing comparison of activity on a given frequency over a period of time. Special types of paper are required for this type of recording.

Two companies, Hogan Faximile and Alfax, provide moist process papers which they promise will give several shades of intensity at fast recording speeds. Alfax paper shown promise of being the most useful, since it works on a principle of ion transfer rather than of electro-chemical change. Since the Alden sales dept. was unable to supply a demonstrator, two sample rolls of Alfax paper were purchased and tested in our lab on a jurry-rigged tape puller. It was found that the moisture content of the paper was critical, as was the electrode material and shape. Without the use of a demonstrator machine, which has been designed to enclose the paper to control the moisture loss, the conclusions of the experiment were limited to determining that about 150 volts at 50 milliamp peaks may be required to record pulse signals.

Spark discharge paper from a Telefax machine was tried, however, it was found that pulse signals were not noticable against the normal pinholes in the paper. This test may be considered inconclusive since there are several other instrument-grade papers available.

b. Camera Recorders

As a result of the studies conducted thus far, it appears that the most expedient way to make permanent records would be to use a cathode ray tube (which may be a memory type) and a Polaroid type camera. To get maximum information on each piece of film, it is possible that CRT's may be multiple-gunned devices or single-gun types grouped together so that a camera can see more than one tube at a time. The suggested camera has the advantage that it can rapidly present the recorded information to the operator permitting him to observe and correct conditions enabling him to make improved photgraphs. The camera makes efficient use of a small amount of paper to record a given amount of useful information. Photographs are small, compact, easily stored and can be reduced to microfilm.

For permanent records the breadboard receiver will utilize a Polaroid type camera which provides clear, rapidly obtained, and easily stored records.

3. Indicator Scanning

For many receiver applications, a coarse indication of signal activity is adequate. For more detailed analysis of the signal activity on each band, it is convenient to examine only a sector of the entire band.

Sector scanning may be advantageous for two purposes. The first is to obtain more resolution over a given range of the receiver sweep. The second is to reduce the acquisition interval by increasing the intercept probability. The operator of the receiver will at times have doubt as to whether an acutal signal exists at a given frequency. By concentrating the scan of the receiver about the frequency at which he believes the signal has been acquired, he may shorten the time required to confirm the actual existence of a true signal. He therefore requires some form of control which allows him to narrow down the range of frequencies which is being swept by a receiver channel. The best form of control would be one in which the operator does not at any time lose the ability to see the frequency at which the investigation will be conducted. The form which the actual control will take was chosen after considering the many methods to be described. Once the sector has been chosen, the best presentation of the sector on an indicator is considered.

The sector may be presented in its true calibrated frequency location on the indicator device as though it were a small piece sliced out of a normal full range sweep or it may be expanded to cover the full width of the indicator device. In the latter case, the provision of accurate frequency calibration of the sector must be considered. In Appendix X various considerations of indicator scan control are discussed, along with an analysis of their effect on intercept probability, acquisition interval, and resolution improvement.

4. Tests Conducted

Figure 8 shows photographs taken on a Hughes Type 106 Memoscope. Figure 8a shows a falling raster presentation. The horizontal axis represents a 4.0 Gc to 7.2 Gc sweep. The signal is visible at approximately 5.6 Gc. The sweep speed was 100 cps and the prf was 600 cps. The vertical sweep was accomplished in 5 seconds. Figure 8b shows the integration over a 1 minute period. Figure 8c shows a 105 prf, and demonstrates how drifting through synchronism affects the detection probability. Figure 8d shows a 3 minute integration of Figure 8c. Figure 9a shows an A-scan as presented on the Memoscope. This represents a single pulse acquisition. Figure 9b demonstrates pulse acquisition in the presence of instantaneous and intentional noise. The signal is unintelligible in this figure. It may, therefore, be concluded that a falling raster presentation on a Memoscope yields the greatest signal detection sensitivity by virtue
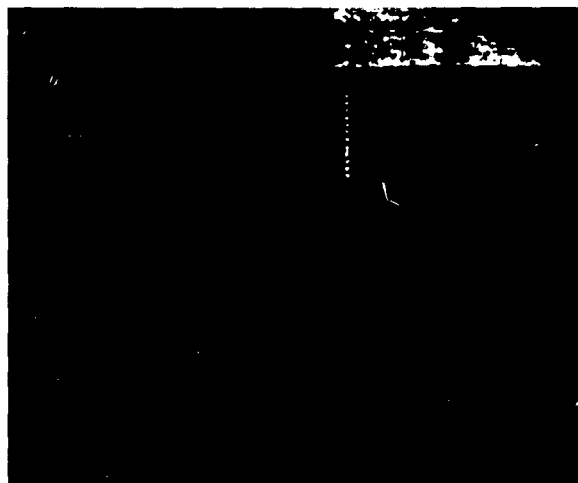
<div align="center">

(a)

Sweep = 100 cps   PRF = 600

5 Second Integration
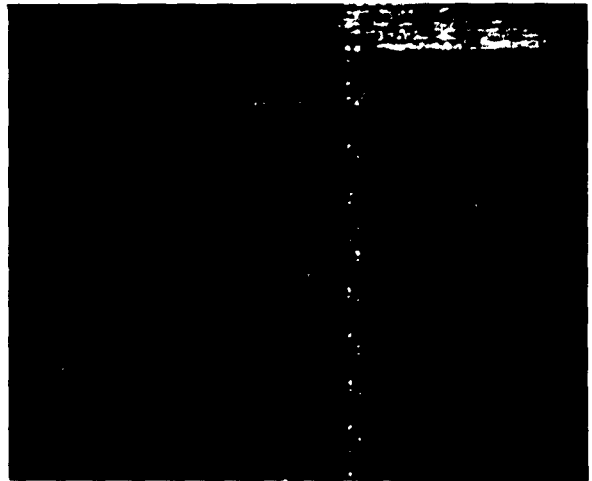
(b)

Sweep = 100 cps PRF = 600
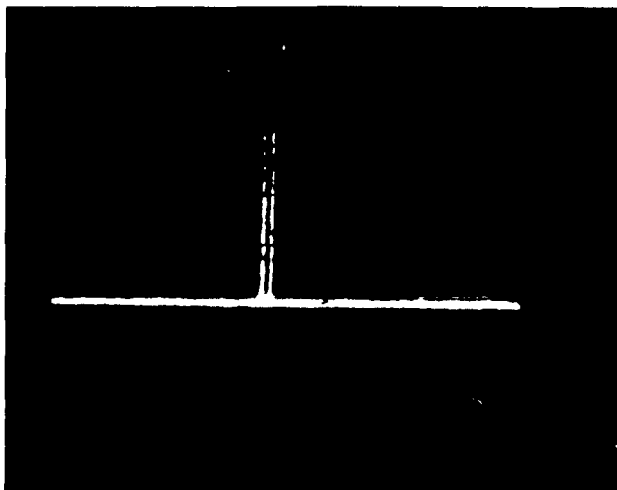
1 Minute Integration

(c)

Sweep = 100 cps   PRF = 105

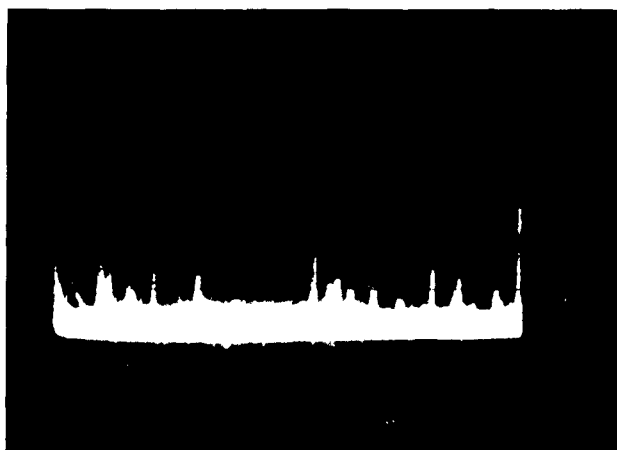15 Second Integration

(d)

Sweep = 100 cps   PRF = 105

3 Minute Integration

Figure 8.   Hughes Type 106 Memoscope Falling Raster Scan

</div>

(a)

Single Pulse Acquisition

(b)

Multiple Pulse Acquisition
With Instantaneous Noise

Figure 9. Hughes Type 106 Memoscope A-Scan

of the signal-to-noise improvement incurred via the integration process. The A-scan is obviously superior for signal analysis.

Figure 10a shows a falling raster scan photographed on a conventional cathode ray tube. In this technique, film is utilized to provide the integration, the camera shutter being held open for extended periods. Figure 10a shows a receiver sweeping the 4.0 Gc to 7.2 Gc range at a rate of 80 sweeps/sec. The signal has a prf of 600. Figure 10b shows the increase of signal intercept when sector scanning is utilized. Here, the receiver is sweeping only 5.75 Gc to 6.25 Gc. Figures 10c and 10d show the effects of a prf which is almost equal to the receiver sweep speed, 80 cps. Figure 11a and 11b show a prf of 40 with a sweep speed of 80 cps.

The effects of visible signal-to-noise improvement via integration can be seen from Figure 11c. Here, a signal with a prf of 600 is sampled by a receiver with a sweep rate of 80 cps. External noise from a 1000 cycle buzzer is added to the circuit. The pulsed signal (as well as the image) is clearly visable.

L. AN/GRR-9 FIELD TEST

A field test was run on the existing AN/GRR-9 receiver in an effort to obtain practical information, data, and typical field problems which might serve in the development of operational and equipment performance concepts under the present program.

Four specific tests were run, each with a specific objective:

Test No. 1

Objective: To determine the effectiveness of an AN/GRR-9 type system in intercepting and observing transmitter harmonic and spurious outputs.

Test No. 2

Objective: To determine the effectiveness of, and the problems resulting from the use of the "Signal-Substitution Method" to measure frequency and signal level on a swept receiver.

Test No. 3

Objective: To investigate the effects of radar pulse repetition rates (PRF) and radar-antenna rotation rates (RPM) on intercept capabilities of a swept receiver as functions of receiver-display sweep rates and cathode-ray tube screen presistency.

Test No. 4

Objective: To observe the operation of the AN/GRR-9 type swept receiver in intercepting and indicating the presence of multiple radar signals.

The test results are presented in Appendix XI. During the course of the testing, several features were found which would be desirable for

(a)
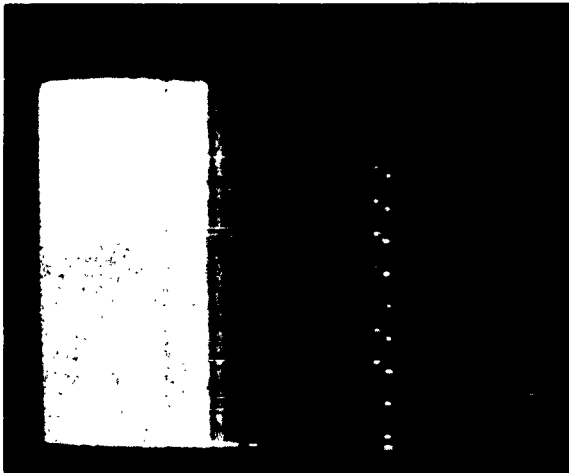
4.0 Gc - 7.2 Gc Range

Sweep = 80 cps   PRF = 600

(b)

5.75 Gc - 6.25 Gc Range

Sweep = 80 cps   PRF = 600

(c)

4.0 Gc - 7.2 Gc Range

Sweep = 80 cps   PRF = 80

(d)

5.75 Gc - 6.25 Gc Range

Sweep = 80 cps   PRF = 80

Figure 10.   Film Integration Falling Raster Scan

<div align="center">

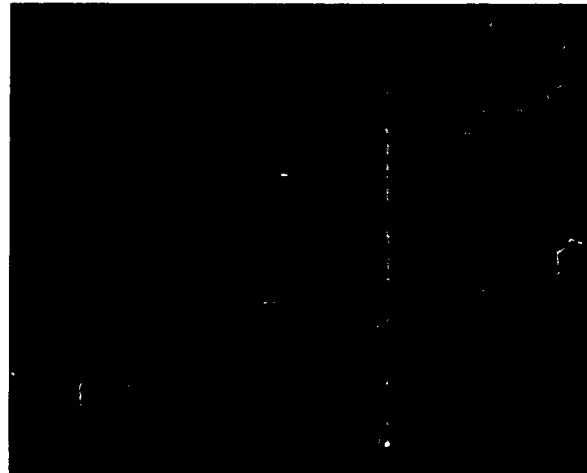(a)
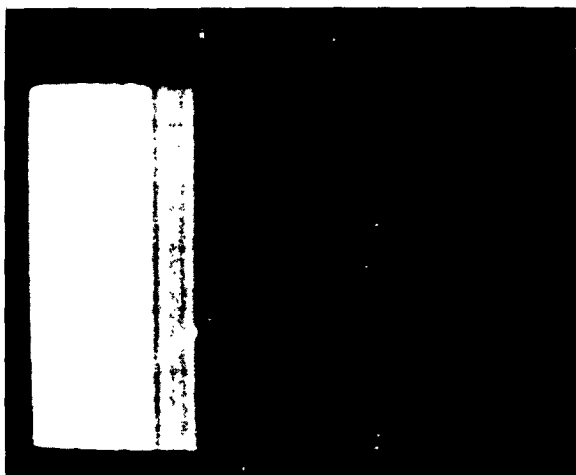
4.0 Gc - 7.2 Gc Range
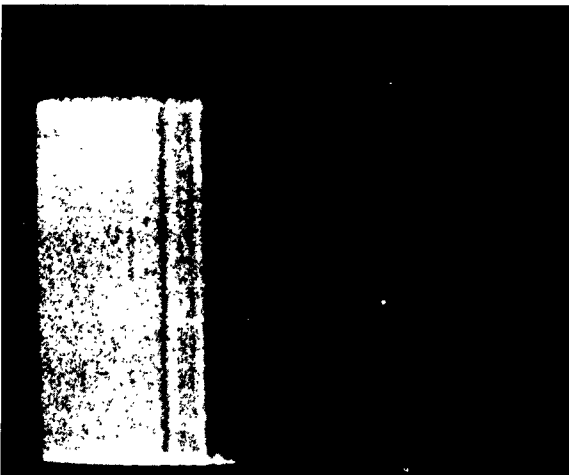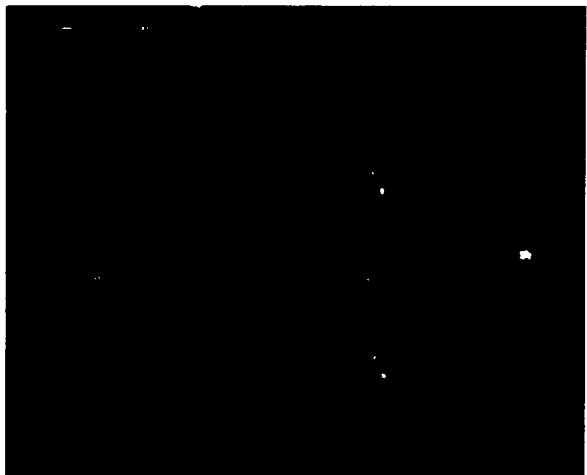
Sweep = 80 cps   PRF = 40

</div>

<div align="center">

(b)
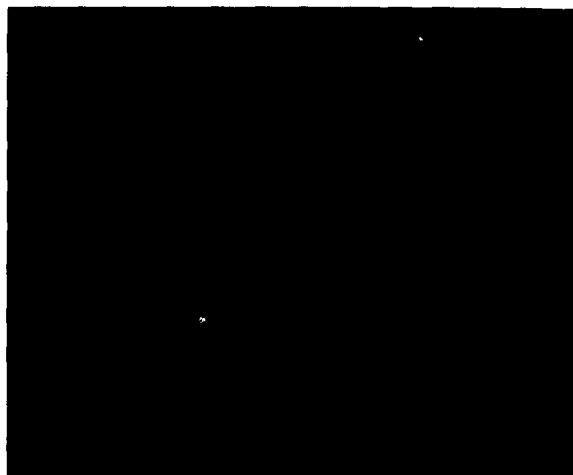
5.75 Gc - 6.25 Gc Range

Sweep = 80 cps   PRF = 40

</div>

<div align="center">

4.0 Gc - 7.2 Gc Range

Sweep = 80 cps   PRF = 600

External Noise Injected

</div>

<div align="center">

Figure 11   Film Integration Falling Raster Scan

</div>

incorporation into a new receiver design:

1. Variable IF bandwidth - to maximize intercept probability and/or increase resolution.

2. Variable sweep width - to maximize intercept probability when the approximate frequency of the signal to be intercepted is known.

3. Paper tape recorder (or other visual recorder) - to record all received pulses. When intercepts are far apart in time, the eye of the observer tires and a blink or a sideways glance may cause a received pulse to be missed. An effort will be made to incorporate all of the above features into the design of the present 100-kc to 40-Gc receivers.

APPENDIX I

RECEIVER COMPARISON

A. TIME-SHARED RECEIVERS

Manually-tuned receivers will obviously not fulfill the requirements imposed by spectrum signature, spectrum surveillance and other types of broadband receivers. The panoramic receiver has evolved as a natural extension of the manually-tuned receiver. Its main features are an automatic frequency sweep and a CRT display designed to yield a signal presentation on a frequency calibrated axis.

In principle, panoramic reception consists of sweeping the frequency band of interest with a bandpass filter. Horizontal deflection of a CRT beam is accomplished synchronously with center frequency sweep. Thus, a direct correspondence exists between the instantaneous center frequency of the filter (or receiver) and the signal position on the indicator display, permitting a calibrated frequency scale to be established.

Both mechanically and electronically tuned panoramic receivers have been utilized. Mechanically tuned receivers are limited to sweep rates on the order of a few cycles per second. Electronically tuned receivers, on the other hand, can achieve sweep rates in the magacycle region. The actual sweep speed required is dependent upon intercept probability, receiver bandwidth and, most importantly, upon the information which the receiver is intended to receive. However, regardless of the sweep speed required, if a time-shared receiver is to be utilized, electronic tuning is favored. Not only does electronic tuning result in a saving in weight and size, but the reliability of an electronically tuned system greatly exceeds that of a mechanically tuned, moving part receiver.

1. Tuned Radio Frequency Receiver

The simplest electronically tuned receiver is shown in Figure I-1. An electronically tuned filter is swept in front of a crystal video receiver. Frequency resolution is dependent upon filter bandwidth. Sensitivity is primarily a function of crystal efficiency, and is normally on the order of -52 dbm for pulsed signals and -65 dbm for CW signals for a 1 Mc video bandwidth (assuming 3.0 db pre-detection loss). If a broadband crystal is utilized, sensitivity is relatively independent of bandwidth.

Electronically tuned filters only suppress spurious responses by a figure on the order of -20 to -40 db. Filter sections can be cascaded to increase this rejection at the expense of increased insertion loss. Moreover, no radiation can exist from the receiving antenna since no oscillators exist within the receiver.

One major disadvantage is that the filter bandwidth cannot easily be changed. Moreover, high Q filters are accompanied by high insertion loss. Thus, the frequency resolution is relatively poor and cannot be changed easily.
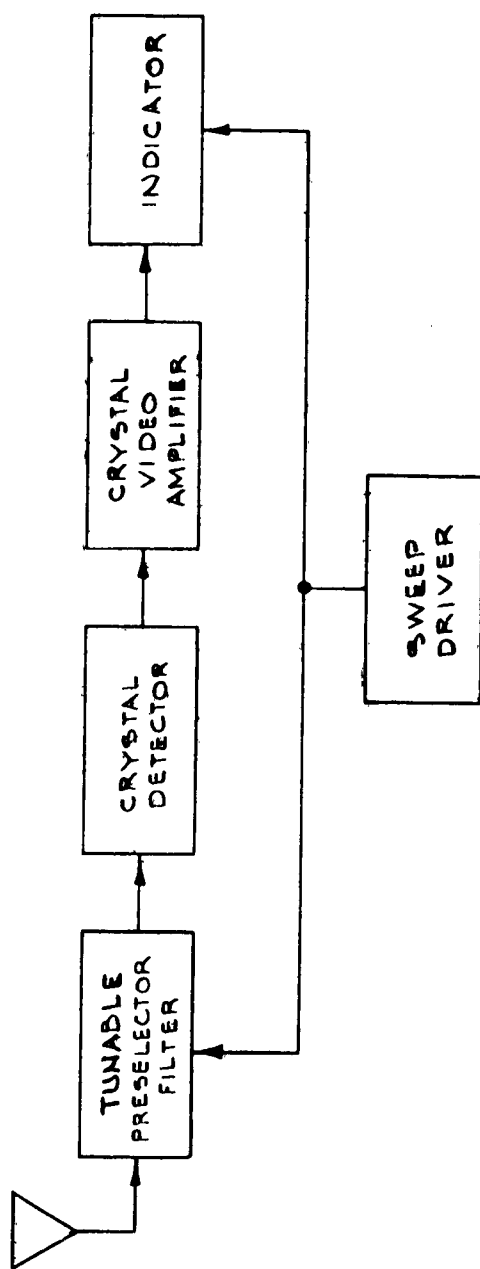
Figure I-1. Tuned Radio Frequency Receiver

1001086

The TRF sensitivities are also low. This can be rectified by the inclusion of a preamplifier ahead of the crystal detector. Traveling wave tube or backward wave tube amplifiers can achieve sensitivities of less than -90 dbm. Moreover, the BWA has a "built in" bandpass filter characteristic. However, the BWA filter is relatively wide so that frequency resolution and spurious rejection are poor.

2. Single Superheterodyne Receiver

Rather than tune a filter ahead of a wide open receiver, the local oscillator of a superheterodyne receiver can be swept in frequency and the I.F. bandpass utilized to provide frequency resolution. A block diagram for a typical voltage-tunable single superheterodyne receiver is shown in Figure I-2.

The superheterodyne receiver has a number of advantages over the TRF receiver. High sensitivities can be obtained since noise figures of 10 db or less are typical. Still lower noise figures can be obtained with low noise preamplifiers. Very high frequency resolution can be achieved with a narrow I.F. bandwidth. Moreover, bandwidth can be easily changed if a selectable bandwidth I.F. is employed.

However, the superheterodyne receiver also gives rise to a number of disadvantages. Additional signals may appear in the I.F. passband due to harmonics of the signal frequency or to conversion products other than the desired mixing products. Moreover, signals outside the passband (e.g. image, I.F. and adjacent-band responses) may cause spurious responses. Spurious signals may also be generated within the receiver local oscillators. If a nonlinear amplifier is employed ahead of the receiver, two or more R.F. signals occurring simultaneously in the amplifier will produce intermodulation responses which may fall within the receiver passband.

Many of these spurious responses can be eliminated by the addition of a tunable bandpass filter preceding the receiver and synchronized to the local oscillator tuning.

Another disadvantage of the superheterodyne receiver is the increased cost over that of a TRF reciever. This is due mainly to the tunable local oscillator required. Backware wave oscillators or voltage tunable magnetrons of $2000 and more commonly utilized for local oscillators. Solid state tunabel oscillators will not only reduce the oscillators cost, size, and weight, but will also decrease the power supply required for driving the local oscillators.

Each receiver band must utilize a separate local oscillator, since, if one local oscillator is used on more than one band, the I.F. frequencies will not only change, but some I.F. amplifiers will necessarily have very high center frequencies.

3. Double Superheterodyne Receiver

A block diagram of a typical double superheterodyne receiver is shown in Figure I-3a. Here the first local oscillator sweeps through the R.F. band, converting incoming signals to a relatively high (usually
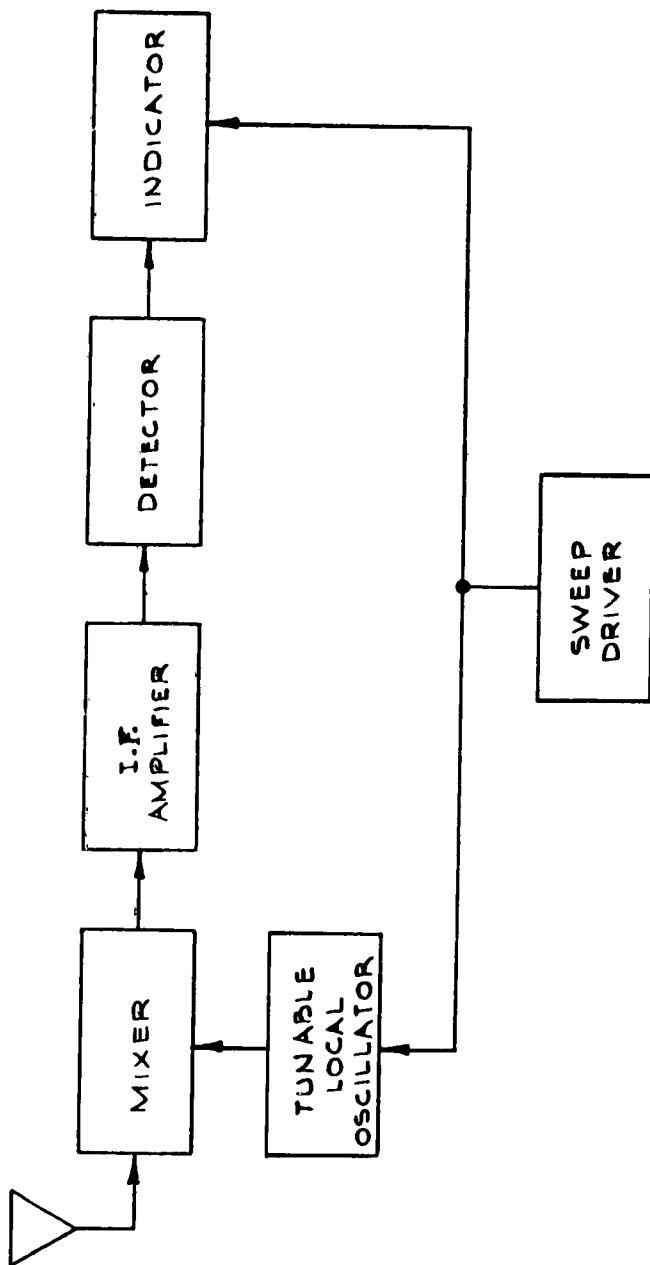
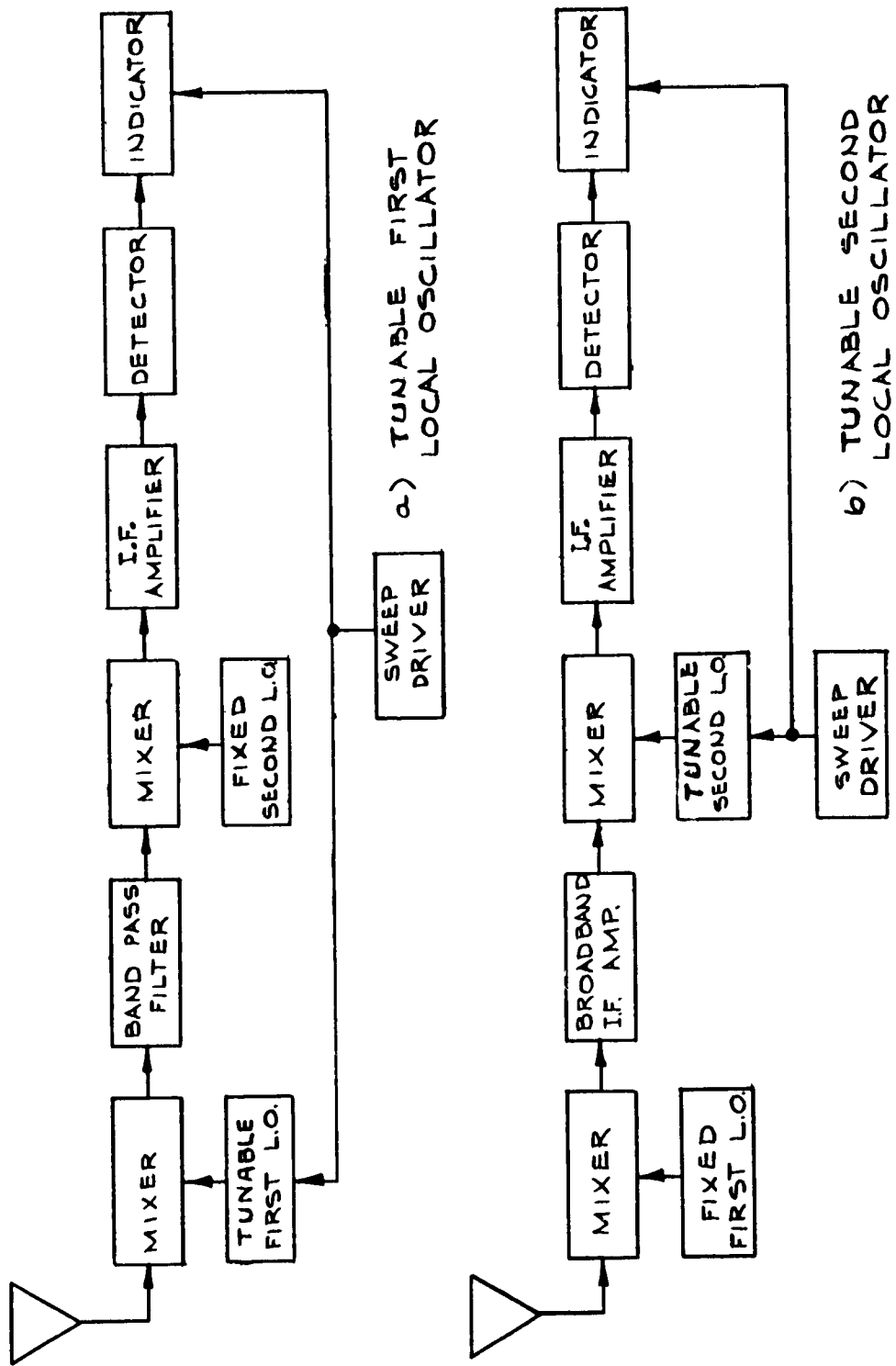Figure I-2. Typical Single Superheterodyne Receiver.

I-4

a) TUNABLE FIRST LOCAL OSCILLATOR

b) TUNABLE SECOND LOCAL OSCILLATOR

Figure I-5. Typical Double Superheterodyne Receiver.

1001097

I-5

microwave) intermediate frequency. The bandpass filter between the two
mixers is used to provide preliminary frequency resolution. The second
local oscillator converts the fixed frequency signals from the filter
to the second I.F. frequency. The bandpass of the second I.F. determines
the receiver frequency resolution.

The possibility of radiation from the receiver antenna is drastically
reduced in the double superheterodyne receiver, due to the local oscil-
lator frequencies being far removed from the R.F. passband. However,
more spurious responses due to intermodulation are now possible in the
dual mixing process. A second result of dual mixing is that the noise
figure is increased by approximately 10 db over single mixing, unless low
noise-high gain preamplification is utilized.

Figure I-3b shows a typical double superheterodyne receiver where
the first local oscillator is fixed and the second is swept. The total
R.F. receiver band is limited in extent in this approach due to the
first I.F. bandwidth available.

One major advantage of the double superheterodyne receiver is that
a single tunable oscillator can be utilized for more than one band as
shown in Figure I-4. However, this ability is no asset to the present
program for three reasons:

   (1) The sweep width per band is limited to that attainable in the
        lowest frequency band, or, if harmonic multiplication is uti-
        lized, the higher frequency bandwidths available pose the
        limitation.
   (2) It is desirable to have complete control over the sweep width
        and sweep frequency of each band.
   (3) The cost of multiple solid state oscillators will be much
        less than the cost of multiple backward wave oscillators,
        magnetrons. and associated components.

4. Cyclotron Resonance Receivers

For many years, the effects of electric magnetic fields on charged
particles in motion have been part of the basic knowledge of electronics.
Recently a considerable amount of research and advanced development has
been expended in applying these principles to obtain a device capable of
selecting and detecting radio frequency signals. Described herein are
both the basic principles and also a device that utilizes these con-
cepts to provide an intercept capability.

The cyclotron resonance tube consists of a cathode for emitting
electrons, a grid for accelerating these electrons, and a collector
for gathering the remainder. Two applied fields are utilized, the
magnetic field being oriented axially towards the collector and the
electric field set in a transverse direction.

Electrons emitted from the cathode are accelerated by a low voltage
on the first grid. The electrons are propelled down the relatively long
drift tube. With the application of a constant magnetic field, the
electrons will tend to spiral about the magnetic flux lines. The
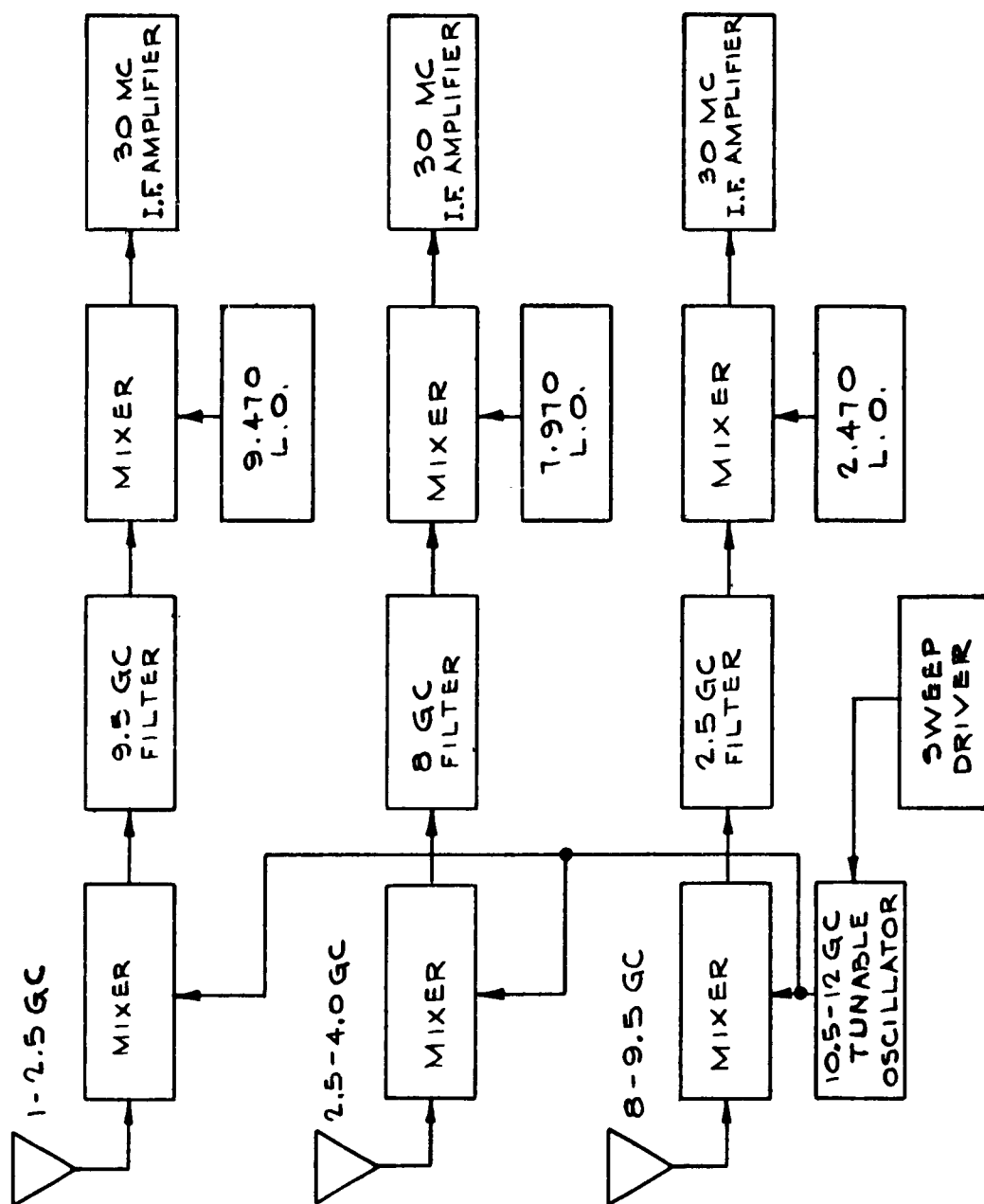
I-6

Figure I-4. Typical Double Superheterodyne Receivers using Common

natural frequency of the spiraling electrons is called the "cyclotron frequency". This frequency, f_0, is related to the magnetic field by the expression, $f_0 = CB$, where C is a constant, and B is the magnetic flux density.

If an R.F. electric field of the same frequency as the cyclotron frequency is applied to the tube, a resonant condition will occur. As a result of this resonance, the radius of the spiral will constantly increase as the electron drifts down the tube. The principle is exactly the same as that which enables particles to gain energy from the electric field in high powered cyclotrons used in nuclear research.

The change in average helical radius can be detected by a honeycomb grid located in the electron path. The interception of electrons by the grid is a function of several factors, one of which is the diameter of the holes in the grid. Resonance, due to the presence of an R.F. signal field, will be indicated by an increase in detector grid current.

Electronic tuning through the use of sawtooth or sinusoidal modulation of the magnetic field is possible. Linear tuning can be achieved since the resonant frequency varies linearly with flux density. For display purposes, a sweeping voltage also can be applied to the Z-axis of an oscilloscope and hence this axis may be calibrated directly in frequency. The Y-axis input of the oscilloscope would be connected to the detector grid. Consequently, as the cyclotron resonance tube is swept through its frequency range, pips will appear on the oscilloscope at positions that correspond to the intercepted frequencies.

An experimental cyclotron resonance receiver has been built at Stanford Electronics Laboratories. Operating in the 65 Mc to 650 Mc band, the tunable receiver provides -90 dbm sensitivity with a low noise amplifier (4 Mc bandwidth). Frequency tuning was found to be linear with solenoid current.

Although the cyclotron resonance tube provides electronic tuning with inherent AM detection with adequate sensitivity (with a pre-amp), no commercial tubes are presently available. Moreover individual tube designs are required for each band. Each band will have an optimum bandwidth which can only be varied slightly.

Sweep speed is limited since a large magnetic field change is required for tuning.

B.   NON-TIME-SHARING RECEIVERS

1.   Crystal Video Receiver

As stated previously, the crystal video receiver is a form of TRF receiver where any signal within a desired R.F. bandpass may be received. A block diagram is shown in Figure I-5.

While the sensitivity of a crystal video receiver is relatively constant with frequency (only varying due to the amount by which the broadband detector is mismatched at a given frequency), the sensitivity is relatively poor. Therefore, for purposes of this comparison, the
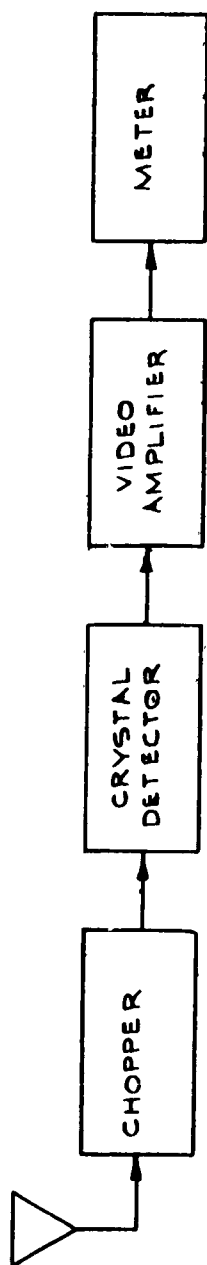
Figure I-5. Crystal Video Receiver.

crystal video receiver will be considered to be preceded by a low-noise preamplifier. This preamplifier will increase the sensitivity in direct proportion to the gain until noise originating within the preamplifier becomes significant. Beyond this point, the characteristics of the video detector also have a bearing on the signal-plus-noise to noise ratio, S + N/N.

Preamplification also increases the receiver dynamic range. Moreover, because of the square law characteristic of the detector, the output range is actually doubled in db relative to the increased input signal range due to the preamplifier.

Although 100% intercept probability is achieved in the crystal video receiver, the lack of frequency discrimination (other than knowing a signal is within the crystal detector frequency response) makes the receiver useless for the present application, except possibly as an alerting function.

2. Sideband Mixing Superheterodyne

The sideband mixing superheterodyne technique endeavers to overcome the frequency resolution limitation of the crystal video receiver while maintaining a sensitive, wide open front end.

A typical block diagram is shown in Figure I-6. Instead of a single L.O. frequency, a series of harmonically related L.O. frequencies is fed to the mixer and allowed to beat with the incoming signal. Since the separation between L.O. frequencies is made equal to the I.F. bandwidth, each sector of the R.F. passband will have an individual L.O. frequency to convert the signal to the I.F. frequency.

The R.F. bandwidth is limited by the harmonic power of the local oscillator, since, as the harmonic power decreases, the conversion efficiency at the harmonic frequency increases with a resultant loss in sensitivity. Moreover, I.F. bandwidth is fixed and resolution is poor relative to a conventional superheterodyne receiver.

Another major disadvantage is low sensitivity. Since the entire R.F. band is open to signal reception, all noise in the R.F. band is converted to the I.F. frequency. Thus, receiver sensitivity is determined by R.F. bandwidth rather than I.F. passband.

3. Discriminator Receiver

The objective of the discriminator receiver is to translate frequency deviations into amplitude variations. A frequency readout is then obtained from the direct reading amplitude analogue.

Figure I-7 is a block diagram of a typical frequency discriminator receiver. The signal from the preamplifier is passed through two networks, one having a flat frequency response (attenuator or directional coupler), while the other exhibits varying frequency response, preferably linear. A pair of matched reversed cyrstal detectors is used, each coupled to a logarithmic video amplifier. The outputs of the log amplifiers are added by a high gain amplifier whose output is fed back
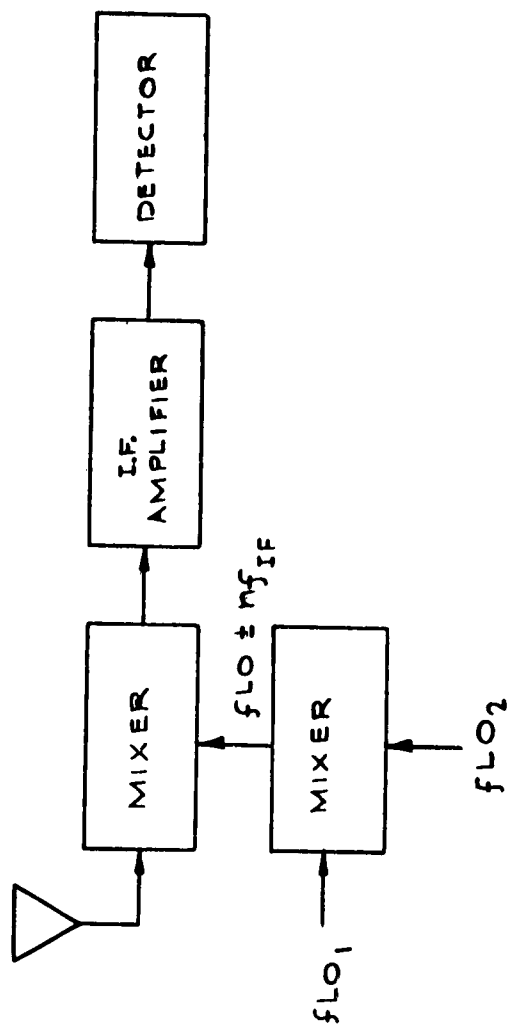
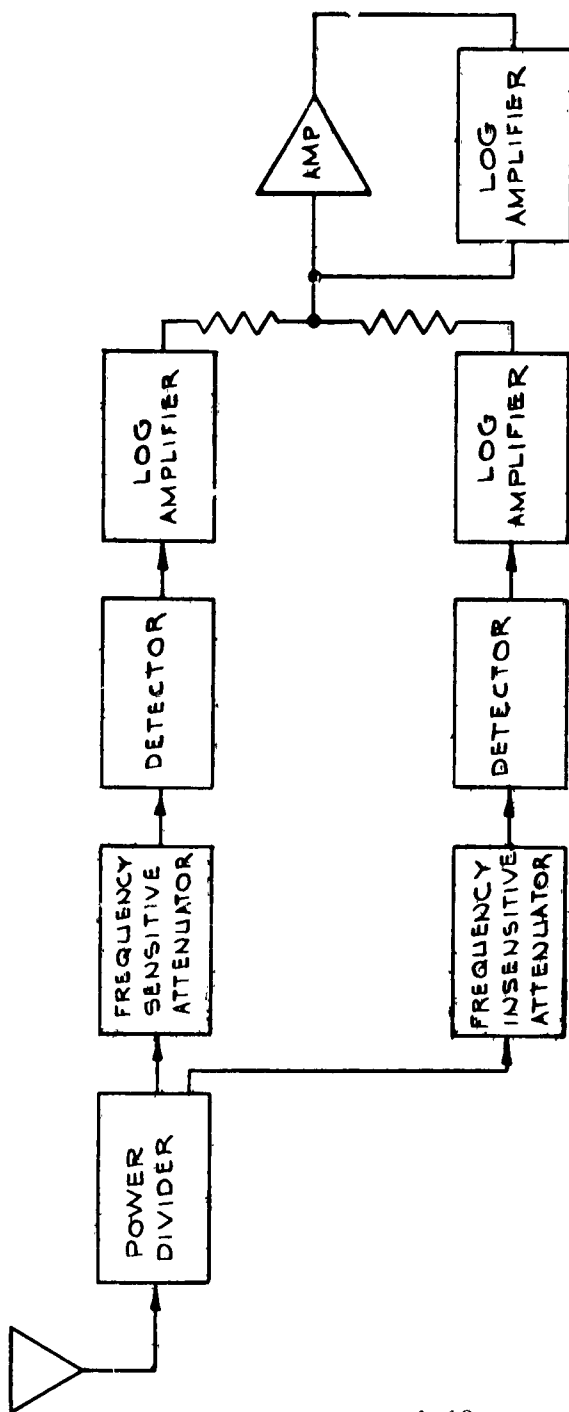Figure I-6. Typical Sideband Mixing Superheterodyne Receiver.

Figure 1-7. Typical Frequency Discriminator Receiver.

100109S

1-12

through a log amplifier to the high gain amplifier input. The output voltage is independent of the input power, but directly dependent upon frequency.

Although offering 100% detection probability together with fair frequency resolution capability, the discriminator receiver requires a relatively high signal to noise ratio for successful operation. Moreover, the discriminator is extremely susceptible to ambiguities due to time coincident signals.

4. **Fully-Channelized Receiver**

The fully-channelized system comes closest of all non-time sharing receivers to satisfying the diverse requirements placed on a universal surveillance receiver. This system is capable of providing the sensitivities and selectivity of a slowly sweeping panoramic receiver and, at the same time, the wide-open feature of a crystal video receiver. Its major disadvantage is the relatively large number of channelizing filters required. The receiver can either be constructed as a superheterodyne or a TRF device.

Figure I-8 is a block diagram of a typical channelized TRF receiver. Figure I-9 shows a typical filter network response curve.

The frequency resolution is completely determined by the selectivity and total bandwidth of the channelized filters.

Figure I-10 is a block diagram of a typical superheterodyne channelized receiver. The only differences from the TRF receiver are: preamplifier gain must now be sufficient to overcome the noise in the mixer; and, a very stable L.O. is required to achieve accurate frequency resolution.

Aside from the disadvantage of numerous filters, special circuitry is required to eliminate ambiguities in frequency at the filter crossover points.
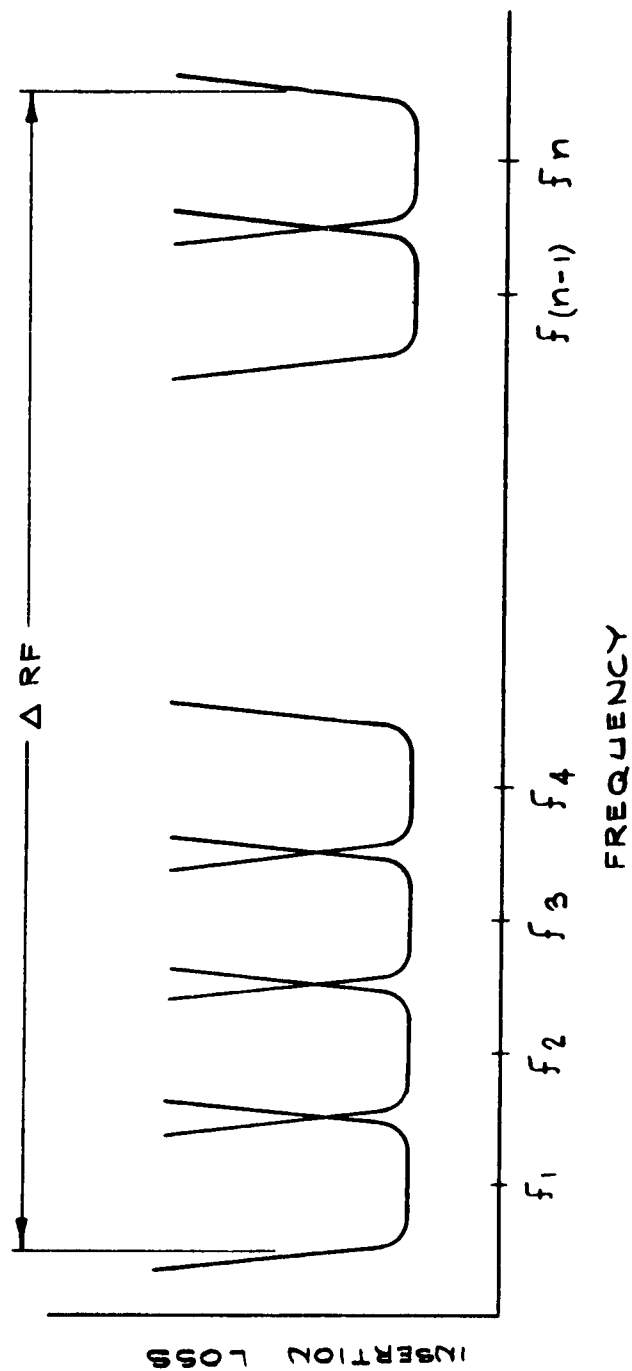
Figure I-9. Typical Filter Network Response Curve for a
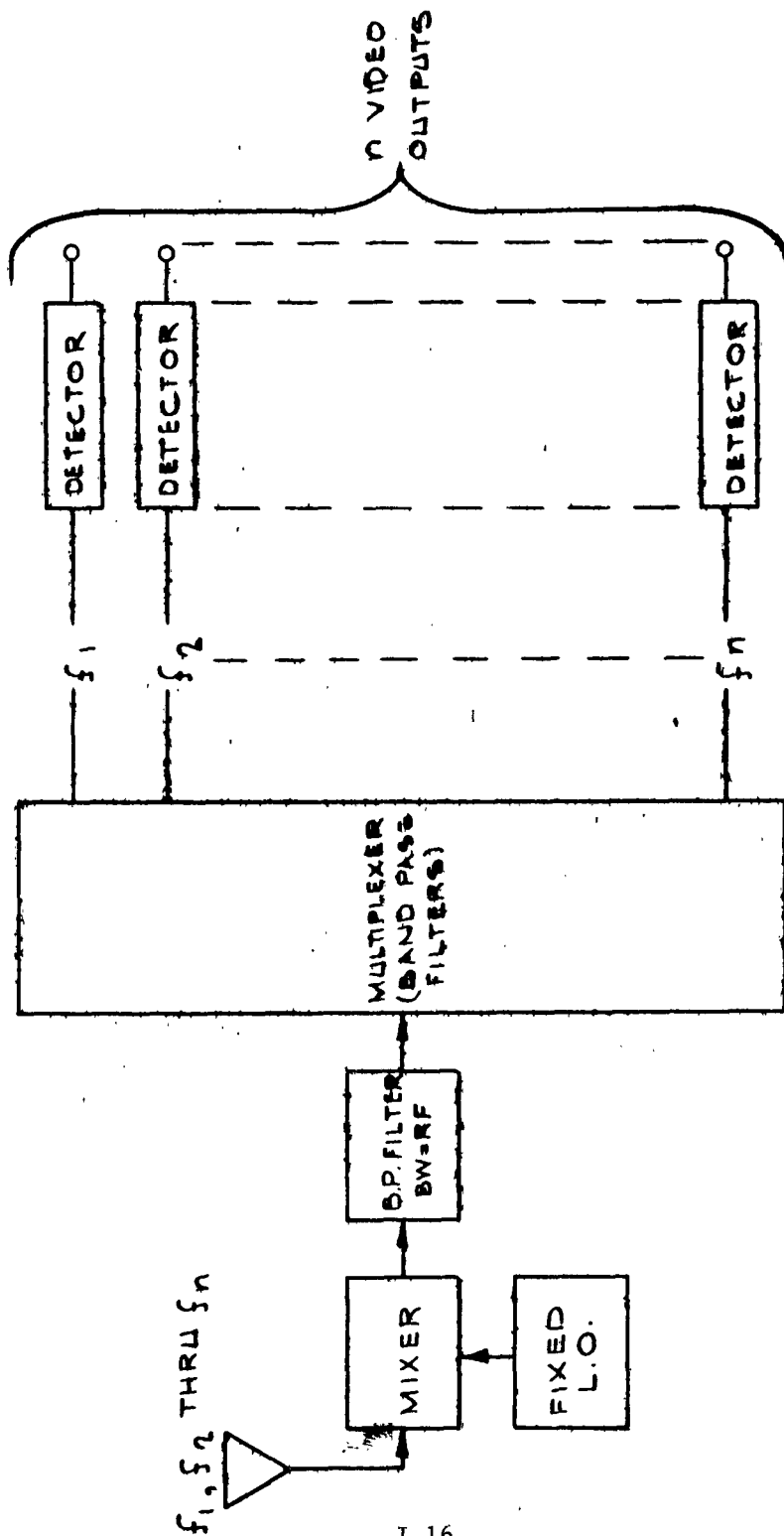Full -Channelized Receiver

1001096

I-15

Figure I 10. Typical Superheterodyne Fully-
Channelized Receiver.

APPENDIX II

BAND SELECTION

Prior to breaking down the 100 Kc to 40 Gc spectrum into bands, it is necessary to study the type of signals and signal densities existing in the total frequency range.

The electronic spectrum is broken down into the following frequency ranges:

    (1) Low frequency (LF): 30 kc - 300 kc
    (2) Medium frequency (MF): 300 kc - 3000 kc
    (3) High frequency (HF): 3 Mc - 30 Mc
    (4) Very high frequency (VHF): 30 Mc - 300 Mc
    (5) Ultra high frequency (UHF): 300 Mc - 3000 Mc
    (6) Super high frequency (SHF): 3 Gc - 30 Gc
    (7) Extremely high frequency (EHF): 30 Gc - 300 Gc

The types of signals and signal activity in each of the above ranges will now be investigated.

From 100 kc to 300 kc, signal types are radio navigation, point to point, and maritime and aeronautical mobile. This portion of the spectrum is not very active since there is no real advantage to be gained in its use: ground wave attenuation is high and equipment is necessarily large.

The 300 kc to 3000 kc band is primarily filled with commercial broadcast signals. Additional commercial uses are for maritime and aeronuatical navigation, amateur, and distress calls. Below 500 kc some military navigation and communications is present, but signal density is low. Between 1.5 Mc and 3.0 Mc, tactical radio nets are present and density is high.

The 3 Mc to 30 Mc band is heavily laden with all types of communication signals, fixed and mobile, on land, sea and air.

Between 30 Mc and 300 Mc, television, FM broadcast, aeronautical navigation, fixed point and mobile transmission and amateur signals crowd the spectrum. Military short distance communication sets also operate in the VHF spectrum.

Commercially the 300 Mc to 3000 Mc range is utilized for fixed, mobile amateur, aeronautical navigation, broadcasting, microwave radio and television relay nets, IFF and radar (including proximity fuses, SHORAN, and missile beacons). Already dense, the UHF spectrum will see even more use in the future.

Commercial and military communication links are present in the 3 Gc to 40 Gc range. More activity is presently planned for the SHF spectrum. Many search and track radars (ground and airborne), and beacons also exist up to 40 Gc.

In addition to adequate resolution capability within a given frequency band, the bandwidth of available components must also be considered. Signal density has been seen to be low in the 100 kc-500 kc

band, however, it is doubtful that the 5:1 frequency range can be easily swept with a narrow band local oscillator. Therefore this band must be divided into two subbands: (1) 100 kc-250 kc; (2) 250 kc-500 kc. Such division will also ease the requirements imposed on a preselector filter.

Above 500 kc, the spectrum is relatively crowded. While not all components are limited to octave bandwidths, many tunable oscillators and filters can only be optimized over octave bandwidths. It is therefore desirable to utilize octave bandwidths up to 1 Gc:

        (3)   500 - 1000 kc
        (4)   1 -   2 Mc
        (5)   2 -   4 Mc
        (6)   4 -   8 Mc
        (7)   8 -   16 Mc
        (8)   16 -   32 Mc
        (9)   32 -   64 Mc
       (10)   64 - 128 Mc
       (11)   125 - 250 Mc
       (12)   250 - 500 Mc
       (13)   500 - 1000 Mc

Above 1 Gc, signal density will allow full utilization of standard waveguide bandwidths, where components are readily available:

       (14)   1 -   2 Gc
       (15)   2 -   4 Gc
       (16)   4 -   8 Gc
       (17)   8 - 12.4 Gc
       (18)   12.4 -   18 Gc
       (19)   18 - 26.5 Gc
       (20)   26.5 -   40 Gc

Thus, there are 20 bands. Increased resolution can be obtained via the variable sweep rate and bandwidth. A large auxiliary presentation will probably be provided for individual band study.

SPURIOUS-SIGNAL SUPPRESSION

A. GENERAL DISCUSSION

Spurious responses can generally be divided into four major categories:

1. Image Signals

The image signal is characterized by the fact that it is displaced (in the direction of the local oscillator frequency) from the wanted signal by twice the intermediate frequency.

2. Signals in the Receiving Stop-Band

Strong, unwanted signals cause harmonic mixing between the input- and local-oscillator signals due to the nonlinearity of the RF amplifier and/or mixer.

3. Intermodulation and Cross Modulation

Two or more incoming signals may beat together to develop an IF response.

4. Intermediate Frequency Within the Signal- or Local-Oscillator Frequency Bands

The receiver can be completely jammed by a signal or local-oscillator frequency which falls within the passband of the IF amplifier. Undesired effects can also occur because of the presence of strong signals on or near the desired frequency and also because of local-oscillator radiation. The cause of each type of spurious response and the methods for elimination or recuction of the response is discussed in detail in the following paragraphs.

B. IMAGE SIGNALS

1. Source of Image Response

If a wide-open receiver is utilized, the intermediate frequency will be developed for signal frequencies which are above or below the local-oscillator frequency by the intermediate frequency:

$$f_{LO} \pm f_{sig} = f_{IF} \qquad \text{(III-1)}$$

As the receiver is swept over a frequency band it is therefore possible to see each incoming signal twice. For example, assume that F_a and F_b are two incoming signals which are separated by twice the IF. Let f_{LO} represent the local oscillator frequency and f_{IF} the intermediate frequency. In this case, responses can be received for each of the following equations:

$$f_a - f_{LO_1} = f_{IF} \qquad \text{(III-2)}$$

$$F_{LO_2} - f_a = f_{IF} \qquad\qquad \text{(III-3)}$$

$$f_b - f_{LO_2} = f_{IF} \qquad\qquad \text{(III-4)}$$

$$f_{LO_3} - f_b = f_{IF} \qquad\qquad \text{(III-5)}$$

These image responses may be eliminated by putting a preselector between the antenna and the receiver input. The preselector utilized should have sufficient selectivity to attenuate the image signal to a level below the minimum sensitivity of the receiver. To minimize the attenuation requirements of the preselector, the IF frequency should obviously be made as high as possible. However, other considerations are involved in the choice of IF, as will be discussed later.

An image signal can also be generated within the mixer itself due to the nonlinearity of the mixing device. These products are of the form:

$$f_{image} = \pm 2f_{LO} - f_{sig} \qquad\qquad \text{(III-6)}$$

$$f_{image} = \pm (f_{sig} - 2f_{IF}), \text{ where } f_{sig} < f_{LO} \qquad\qquad \text{(III-7)}$$

$$f_{image} = \pm (f_{sig} + 2f_{IF}), \text{ where } f_{LO} > f_{sig} \qquad\qquad \text{(III-8)}$$

Normally, the products developed from Equations (III-7) and (III-8) can be neglected since their amplitued is extremely low. The products formed by equation (III-6) however, will normally be seen by the receiver, but this product will coincide with the true signal, and will therefore cause no spurious-response difficulties if preselection is utilized.

   2. Elimination of Image Response

      a. Preselection

      The image response can be eliminated by the use of preselection in the antenna line, prior to the mixer stage. It is recommended that a bandpass filter be used to protect the overall bandwidth of the front-end while a swept preselector be used to protect the receiver from spurious responses within the band. The preselector should have enough selectivity to attenuate the image signal to the point at which a response on the display is not discernable. To aid in this, the IF will be made as large as possible, keeping in mind that there are other considerations in the choice of the IF frequency. This will aid in the attenuation of the image signal, since the selectivity falls off at a rate of 20 db per decade per tuned pair.

b. Outphasing (Image Rejection Mixer).

A method which can be used for the rejection of the image response
generated within the mixer is the outphasing technique (Figure III-1).
This method is similar to the phase cancellation used in single-sideband
radio. One of the two input signals is shifted 90° in phase and then
the signals are fed to two mixers. The local-oscillator signals which
are fed to the mixers are identically phased. The IF output of one
mixer lags the other mixer output by 90° when the local-oscillator fre-
quency is lower than the signal frequency and leads the other mixer
output by 90° when the local-oscillator frequency is higher than the
signal frequency. An additional 90° phase lag is added to one IF
preamplifier and then the two IF outputs from the preamplifiers are
added. These signals reinforce each other when the local-oscillator
frequency is above the signal frequency because they have equal phase
and they cancel when the local-oscillator frequency is below the signal
frequency (image). Approximately 20 db of image rejection can be
achieved by the use of this technique. The phase cancellation method
gives rise to problems when it is to be used over a wide bandwidth,
such as an octave bandwidth. One problem is that it is difficult to
get a 90° phase shift in one of the signal lines over an octave band;
another is that the circuitry in the two signal lines before the
addition of the two IF signals will have to be phase equal, excluding
the 180° phase shift that is added.

c. Image Blanking

A block diagram of a generic approach to reducing the image re-
jection problem is shown in Figure III-2. This receiver splits the
input power and feeds two channels. By using a frequency translator
or a tracking local oscillator, channel 2 can be driven at a different
frequency. If the frequency difference of the two local oscillators is
twice the intermediate frequency, where the two IF's are identical, one
local oscillator tunes the desired signal from below the signal fre-
quency, while the other tunes from above. It is obvious that the
desired signal is present in both channels whereas the image is present
only in one. This is pictorially represented in Figure III-3. Study
will show that, as a consequence, it is only necessary to compare levels
or correlate the two channels to see if a signal exists at the image
frequency. The output is gated on only when the desired response is
present in both channels. This is particularly useful in swept or
automated systems where false indications are more apparent. Now con-
sider the interference aspect of the image frequency - if the image
signal is coincident with the desired signal, it is only necessary to
identify the channel in which the interference is absent and switch to
that channel. This could also be done electronically. There is still
the question of interference in the opposite channel if a signal also
appears at the image frequency of that channel. If a different inter-
mediate frequency were used in one channel, the image frequency for
that channel would also change. Thus, if signal densities are sufficient
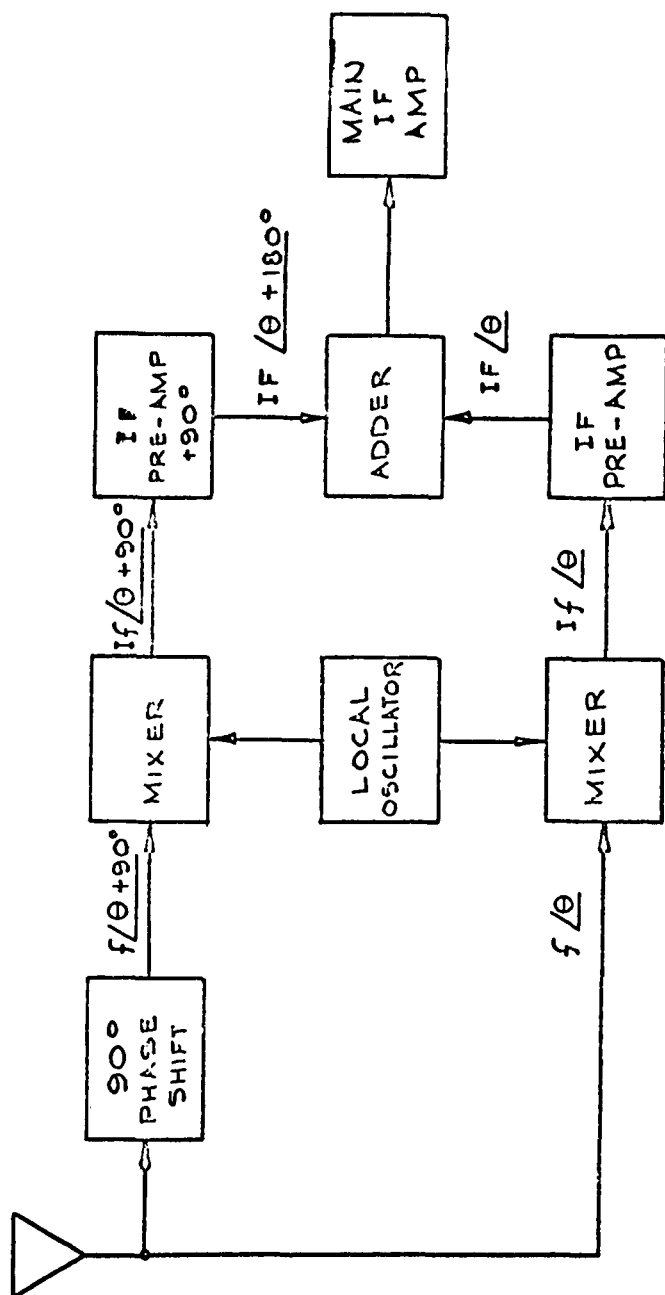to present image interference in both channels and to negate the

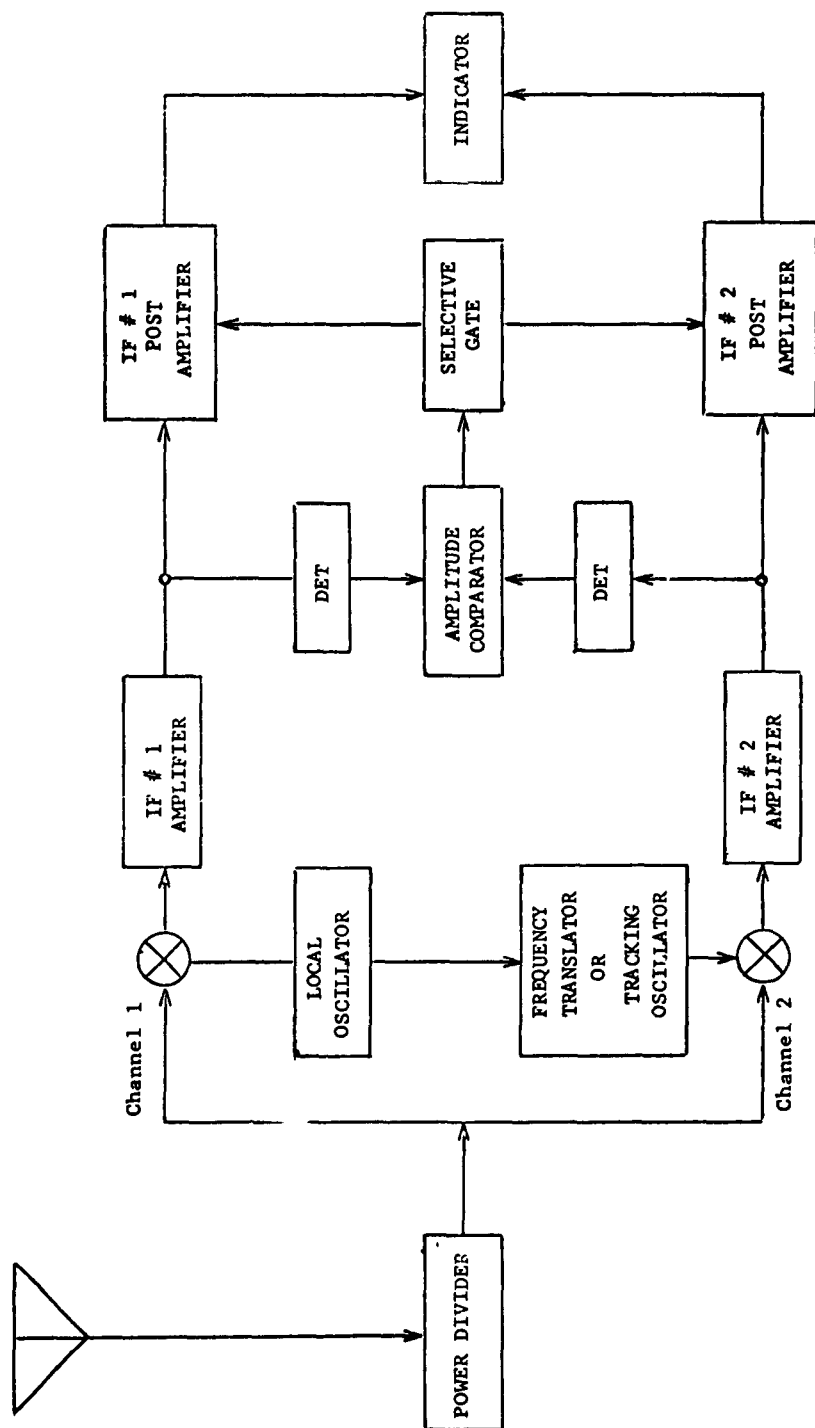Figure III-1. Block Diagram of Image Outphasing Technique

Figure III-2. A Selective Signal Receiver for Image Blanking

effectiveness of choice, it is then possible to provide a tunable IF in one channel in order to eliminate the interference.
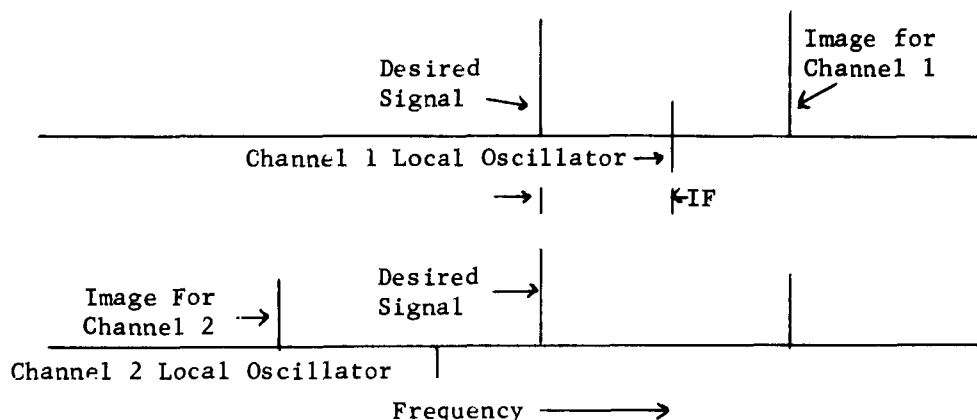

Figure III-3. Pictorial Representation of Frequency Acquisition in an Image-Blanking Receiver

The effectiveness of this system for discriminating against spurious responses has not been determined. Because of the harmonic relationship of the local-oscillator frequency and the signal frequency, some validity is attached to the approach. Such an analysis will be a study requirement before implementing the system.

It is most worthy of mention that the local oscillator of the "unused" channel can provide a very fine source for calibrating the opposite channel.

This image-blanking tehcnique is held to be especially applicable for receivers operating in a measurement mode. The approach is dependent upon finding a vacant frequency band (corresponding to one IF bandwidth) corresponding to the image frequency of one channel. Under such conditions the system can completely eliminate the image response.

C. SIGNALS IN THE RECEIVER STOP-BAND

 1. Source of Spurious Responses

If unwanted signals are defined as those frequencies incident within the stop-band of a receiver, discrimination against these signals can be achieved to the degree permitted by the selectivity characterisitc of the IF amplifier alone. If the stop-band signals are of sufficient amplitude, the nonlinearity of the RF amplifier, in particular, or of the mixer, will result in harmonic mixing between input signals themselves or input signals and local-oscillator signals. When the difference frequency of any ordered combination falls within the IF bandpass, a spurious response occurs. These signals follow in accordance with the well known expression $mf_s \pm nf_{LO} = f_{IF}$ where m and

n are the harmonic numbers of the two mixing signals. When m and n are both unity, the desired response and the image response are the resultants. Further study of the expression will show that the frequencies of the signals producing spurious responses are a function of the intermediate frequency. It follows then, that in a system employing preselection to minimize stop-band signal incidence, the RF selectivity can be eased by a proper selection of the intermediate frequency and consideration of the impinging signal density. In such a system, therefore, the IF should be chosen so that only frequencies far removed from the center frequency could cause a spurious response in the absence of preselection. The preselection and tracking requirements can then be much eased.

This class of spurious response is graphically displayed in Figure III-4, a plot of the spurious responses occurring over the 1- to 2-Gc band when a 30-Mc IF is used. This is a plot of the local oscillator frequency vs. the signal frequency for harmonic orders through 9(m + n = 9).

Consider the example of a 1200-Mc signal for which the normal local-oscillator frequency is 1230 Mc. Assume a receiver with no preselection, and enter the chart at an ordinate value (signal frequency) of 1200 Mc. Maintain the constant 1200-Mc ordinate and move across the chart for various values of local-oscillator frequency. Each line which is crossed corresponds to an output response at 30 Mc when the local-oscillator value is the abscissa of that crossing. It is thus obvious that many local-oscillator values can create and IF output if the output signal is sufficiently large. The amplitude of the various responses diminishes with the harmonic order. Such order is given by the harmonic numbers m + n where m and n correspond to the local oscillator and signal frequencies respectively. The digits (1) and (2) represent real and image responses respectively. The detriment of these particular responses is principally that of falsely indicating several signals in the presence of only one input for which the desired response is 11 (1).

The second detrimental effect of spurious responses is that of interference. By extending the above example to permit any signal input, and by restricting the local-oscillator frequency to a fixed value, e.g., 1230 Mc, the effects are readily noted. Consider a vertical line through 1230 Mc for which the desired input is 1200 Mc. Any spurious line crossing the vertical shows that an input signal corresponding to that particular ordinate will cause a response. Thus, the effect is that of multiple-tuned signals, and interference is involked at a level commensurate with the harmonic order, as previously discussed. Further tuning of the local oscillator, i.e., movement of the vertical line, shows the shift of spurious responses.

The advantage of preselection to negate the effects of false signals and interference is now obvious. An ideal preselector would permit only signals in its narrow band to impinge on the mixer. The

Figure III-4. Plot of L Band Spurious Responses with a 30 Mc IF

III-8

preselector may be shown on the chart by drawing parallel lines on each side of 11(1) at the appropriate width.

1. Elimination of Spurious Responses

   a. Preselection

As has been demonstrated previously (refer to Figure III-4), preselection will reduce greatly the level of image and spurious responses due to signals in the receiver stop-band. A spurious signal at a relative output level of -40 db can be reduced by the amount of the filter skirt rejection of the input signal producing the response. Thus, if the signal is far removed from the preselector center frequency, the response can be eliminated. This justifies the extreme requirements normally imposed on the preselector characteristics as well as the desirability of a high IF frequency.

   b. Choice of IF Frequency

An "optimum" intermediate frequency may be chosen for any system after an appropriate consideration of the requirements. Figure III-4 represents an analysis for a single IF and is very useful for a specific receiver or for tutorial work. The choice of the IF becomes tedious if such an approach is consistently followed. A less obvious but more universal approach is permitted by normalizing the expression $mf_s \pm nf_{LO} = f_{if}$ to $mB \pm nA = f_{IF}$, where B is always considered the higher mixing frequency. A plot of the normalized equation is shown in Figure III-5. The resulting spurious responses are shown in Figure III-6 for an IF of B-A, and in Figure III-7 for an IF of B + A.

Only the frequencies of harmonic-producing sources have thus far been considered, and little attention has been paid to relative amplitudes of spurious responses. In a practical system, amplitude determination is a less exact science but is nonetheless important in establishing the intermediate frequency and bandpass requirements. The following discussion will therefore pertain to spurious-signal amplitude reduction and some supporting data will be offered.

Of the spurious products to be encountered, the image response 11(2) in Figure III-4 has the greatest amplitude, and is generally equal to the desired response. The output responses of higher-ordered products which are relative to the desired signal vary inversely with the order number. The actual magnitude depends upon the nonlinear characteristic of the mixing or modulating device. Where a large system dynamic range is involved, the higher orders of spurious signals are detectable.

Figure III-8 shows the relative levels of spurious responses generated in a 900-Mc mixer with a 30-Mc output. Note that the levels of the 8th order ($4 F_{LO} - 4F_s$) and 10th order ($5F_{LO} - 5F_s$) responses are grouped very closely with greater than -10 dbm signal input. However, about 30 db of additional spurious-signal-free dynamic range is achieved with a 10th order response and no 8th order response. Lower order responses (12th, 14th, etc.) group so closely about the 10th order, that no appreciable spurious-free range is achieved in further harmonic
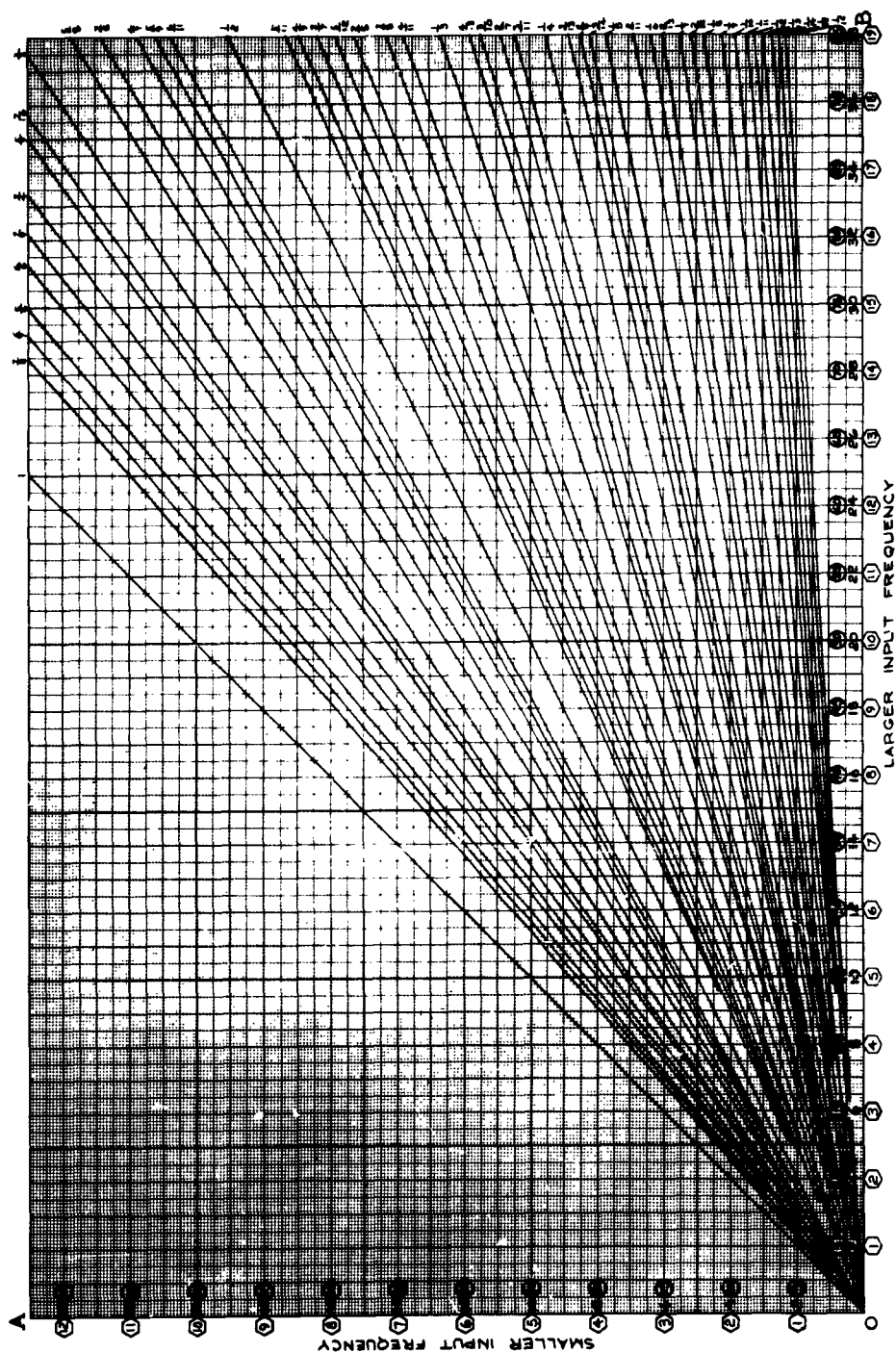
Figure III-5. Plot of Normalized Equation

R = A/B	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
	B	B±A	B±2A	B±3A	B±4A	B±5A	B±6A	B±7A	B±8A	B±9A	B±10A	B±11A	B±12A	B±13A	B±14A
1/16	-	-	-	-	-	-	-	-	-	-	-	-	-	-	15A
1/15	-	-	-	-	-	-	-	-	-	-	-	-	-	14A	-
1/14	-	-	-	-	-	-	-	-	-	-	-	-	13A	-	-
1/13	-	-	-	-	-	-	-	-	-	-	-	12A	-	-	-
1/12	-	-	-	-	-	-	-	-	-	-	11A	-	-	-	2B-13A
1/11	-	-	-	-	-	-	-	-	-	10A	-	-	-	2B-12A	-
1/10	-	-	-	-	-	-	-	-	9A	-	-	-	2B-11A	-	-
1/9	-	-	-	-	-	-	-	8A	-	-	-	2B-10A	-	-	-
1/8	-	-	-	-	-	-	7A	-	-	-	2B-9A	-	-	-	-
2/15	-	-	-	-	-	-	-	-	-	-	-	-	-	-	14A-B
1/7	-	-	-	-	-	6A	-	-	-	2B-8A	-	-	-	13A-B	-
2/13	-	-	-	-	-	-	-	-	-	-	-	-	12A-B	-	-
1/6	-	-	-	-	5A	-	-	-	2B-7A	-	-	11A-B	-	-	-
2/11	-	-	-	-	-	-	-	-	-	-	10A-B	-	-	-	3B-12A
1/5	-	-	-	4A	-	-	-	2B-6A	-	9A-B	-	-	-	3B-11A	-
3/14	-	-	-	-	-	-	-	-	-	-	-	-	-	-	13A-2B
2/9	-	-	-	-	-	-	-	-	8A-B	-	-	-	3B-10A	-	-
3/13	-	-	-	-	-	-	-	-	-	-	-	-	-	12A-2B	-
1/4	-	-	3A	-	-	-	2B-5A	7A-B	-	-	-	3B-9A	11A-2B	-	-
3/11	-	-	-	-	-	-	-	-	-	-	-	10A-2B	-	-	-
2/7	-	-	-	-	-	-	6A-B	-	-	-	3B-8A	-	-	-	-
3/10	-	-	-	-	-	-	-	-	-	-	9A-2B	-	-	-	4B-11A
4/13	-	-	-	-	-	-	-	-	-	-	-	-	-	-	12A-3B
1/3		2A				2B-4A / 5A-B				3B-7A / 8A-2B				4B-10A / 11A-3B	
4/11	-	-	-	-	-	-	-	-	-	-	-	-	10A-3B	-	-
3/8	-	-	-	-	-	-	-	-	-	7A-2B	-	-	-	4B-9A	-
2/5	-	-	-	-	4A-B	-	-	-	3B-6A	-	-	9A-3B	-	-	-
5/12	-	-	-	-	-	-	-	-	-	-	-	-	-	-	11A-4B
3/7	-	-	-	-	-	-	-	-	6A-2B	-	-	-	4B-8A	-	-
4/9	-	-	-	-	-	-	-	-	-	-	8A-3B	-	-	-	5B-10A
5/11	-	-	-	-	-	-	-	-	-	-	-	-	10A-4B	-	-
1/2	A	-	-	3A-B	2B-3A	-	5A-2B	3B-5A	-	7A-3B	4B-7A	-	9A-4B	5B-9A	-
6/11	-	-	-	-	-	-	-	-	-	-	-	-	-	-	10A-5B
5/9	-	-	-	-	-	-	-	-	-	-	-	8A-4B	-	-	-
4/7	-	-	-	-	-	-	-	-	6A-3B	-	-	-	5B-8A	-	-
3/5	-	-	-	-	-	4A-2B	-	-	-	4B-6A	-	-	-	9A-5B	-
5/8	-	-	-	-	-	-	-	-	-	-	7A-4B	-	-	-	6B-9A
2/3	-	-	2A-B	-	-	-	3B-4A	5A-3B	-	-	-	5B-7A	8A-5B	-	-
7/10	-	-	-	-	-	-	-	-	-	-	-	-	-	-	9A-6B
5/7	-	-	-	-	-	-	-	-	6A-4B	-	-	-	6B-8A	-	-
3/4	-	-	-	-	3A-2B	-	-	-	4B-5A	-	-	7A-5B	-	-	-
7/9	-	-	-	-	-	-	-	-	-	-	-	-	-	8A-6B	-
4/5	-	-	-	-	-	-	4A-3B		-	-	5B-6A	-	-	-	-
5/6	-	-	-	-	-	-	-	-	5A-4B	-	-	-	6B-7A	-	-
6/7	-	-	-	-	-	-	-	-	-	-	6A-5B	-	-	-	7B-8A
7/8	-	-	-	-	-	-	-	-	-	-	-	-	7A-6B	-	-
1	-	-	-	2B-2A	-	3B-3A	-	4B-4A	-	5B-5A	-	6B-6A	-	7B-7A	-

Figure III-6.   Spurious Response Chart

R–A/B	ORDER														
	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
0	R	B±A	B±2A	B±3A	B±4A	B±5A	B±6A	B±7A	B±8A	B±9A	B±10A	B±11A	B±12A	B±13A	B±14A
1/14	-	-	-	-	-	-	-	-	-	-	-	-	-	-	2B-13A 15A
1/13	-	-	-	-	-	-	-	-	-	-	-	-	-	2B-12A 14A	-
1/12	-	-	-	-	-	-	-	-	-	-	-	-	2B-11A 13A	-	-
1/11	-	-	-	-	-	-	-	-	-	-	-	2B-10A 12A	-	-	-
1/10	-	-	-	-	-	-	-	-	-	-	2B-9A 11A	-	-	-	-
1/9	-	-	-	-	-	-	-	-	-	2B-8A 10A	-	-	-	-	-
1/8	-	-	-	-	-	-	-	-	2B-7A 9A	-	-	-	-	-	-
1/7	-	-	-	-	-	-	-	2B-6A 8A	-	-	-	-	-	-	-
2/13	-	-	-	-	-	-	-	-	-	-	-	-	-	-	3B-12A 14A-B
1/6	-	-	-	-	-	-	2B-5A 7A	-	-	-	-	-	-	3B-11A 13A-B	-
2/11	-	-	-	-	-	-	-	-	-	-	-	-	3B-10A 12A-B		-
1/5	-	-	-	-	-	2B-4A 6A	-	-	-	-	-	3B-9A 11A-B	-	-	-
2/9	-	-	-	-	-	-	-	-	-	-	3B-8A 10A-B	-	-	-	-
1/4	-	-	-	-	2B-3A 5A	-	-	-	-	3B-7A 9A-B	-	-	-	-	4B-11A 13A-2B
3/11	-	-	-	-	-	-	-	-	-	-	-	-	-	4B-10A 12A-2B	-
2/7	-	-	-	-	-	-	-	-	3B-6A 8A-B	-	-	-	-	-	-
3/10	-	-	-	-	-	-	-	-	-	-	-	-	4B-9A 11A-2B	-	-
1/3	-	-	-	2B-2A 4A	-	-	-	3B-5A 7A-B	-	-	-	4B-8A 10A-2B	-	-	-
4/11	-	-	-	-	-	-	-	-	-	-	-	-	-	-	5B-10A 12A-3B
3/8	-	-	-	-	-	-	-	-	-	-	4B-7A 9A-2B	-	-	-	-
2/5	-	-	-	-	-	-	3B-4A 6A-B	-	-	-	-	-	-	5B-9A 11A-3B	-
3/7	-	-	-	-	-	-	-	-	-	4B-6A 8A-2B	-	-	-	-	-
4/9	-	-	-	-	-	-	-	-	-	-	-	-	5B-8A 10A-3B	-	-
1/2	-	-	2B-A 3A	-	-	3B-3A 5A-B	-	-	4B-5A 7A-2B	-	-	5B-7A 9A-3B	-	-	6B-9A 11A-4B
5/9	-	-	-	-	-	-	-	-	-	-	-	-	-	6B-8A 10A-4B	-
4/7	-	-	-	-	-	-	-	-	-	-	5B-6A 8A-3B	-	-	-	-
3/5	-	-	-	-	-	-	-	4B-4A 6A-2B	-	-	-	-	-	-	-
5/8	-	-	-	-	-	-	-	-	-	-	-	-	6B-7A 9A-4B	-	-
2/3	-	-	-	-	3B-2A 4A-B	-	-	-	-	5B-5A 7A-3B	-	-	-	-	1B-8A 10A-5B
5/7	-	-	-	-	-	-	-	-	-	-	-	6B-6A 8A-4B	-	-	-
3/4	-	-	-	-	-	-	4B-3A 5A-2B	-	-	-	-	-	-	7B-7A 9A-5B	-
4/5	-	-	-	-	-	-	-	-	5B-4A 6A-3B	-	-	-	-	-	-
5/6	-	-	-	-	-	-	-	-	-	-	6B-5A 7A-4B	-	-	-	-
6/7	-	-	-	-	-	-	-	-	-	-	-	-	7B-6A 8A-5B	-	-
7/8	-	-	-	-	-	-	-	-	-	-	-	-	-	-	8B-7A 9A-6B
1	-	2B 2A	-	3B-A 3A-B	-	4B-2A 4A-2B	-	5B-3A 5A-3B	-	6B-4A 6A-4B	-	7B-5A 7A-5B	-	8B-6A 8A-6B	-

Figure III-7.. Spurious Response Chart

I_{XTAL} = 1.0 MA
P_{LO} = .45 MW, -3.5 DBM
F_{LO} = 900 Mc

MIXER OUTPUT IN DBM

III-13
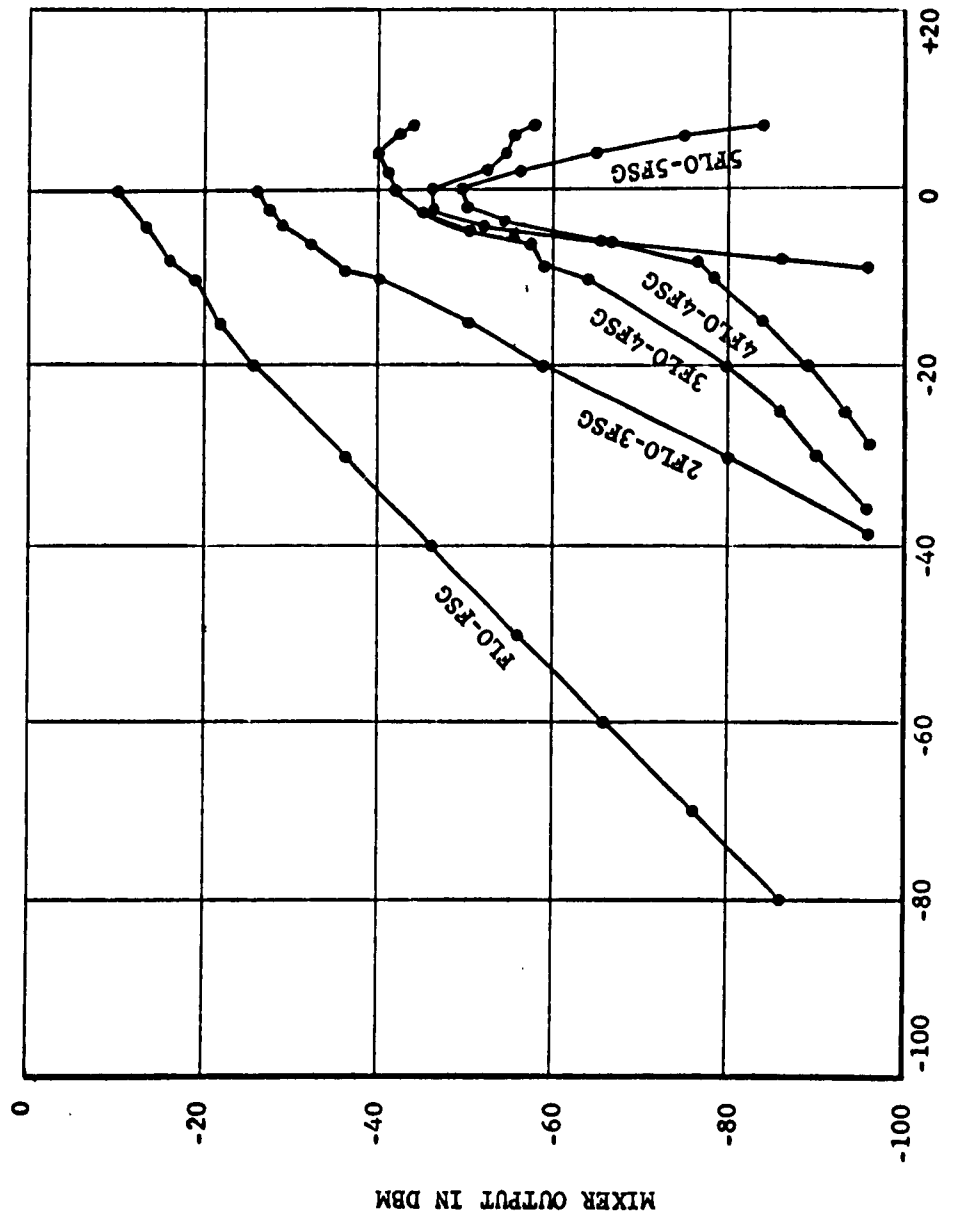
order suppression. Therefore, responses of greater than the 10th order will be allowed. Values of local oscillator and IF frequencies which will allow only spurious responses below the 10th order are given in Table III, of the text.

Assuming a receiver sensitivity of -90 dbm, an 80-db dynamic range free of spurious signals can be achieved, as shown in Figure III-8. Greater than 90 db of spurious-response suppression will exist at signal input levels below -10 dbm (throughout the receiver dynamic range).

   c. Balanced Mixer to Minimize Spurious Responses.

By properly combining the outputs from a balanced mixer, certain spurious responses can be attenuated by approximately 50 db. Figures III-9 and III-10 indicate which spurious products may be attenuated by the utilization of a balanced mixer configuration. All theoretical calculations are performed on the assumption that single-ended mixers are used. Balanced mixers will supply a greater "safety factor".

   d. Local-Oscillator Drive Power to Minimize Spurious Responses

In a system which uses a crystal mixer, some reduction of the level of spurious signals can be afforded by increased local-oscillator drive, which allows for more linear operation. Figures III-8 and III-11 show the effects of an increased local-oscillator drive power at 900 mc for a 30 Mc IF. The most significant harmonic responses are plotted. At the normal crystal current level of 1 ma, approximately 50 db of dynamic range free of spurious signals is provided (assuming all spurious products are allowed to be generated). If the drive is increased to 30 ma, a 60-db dynamic range free of spurious signals results. All other responses follow this pattern. As might have been anticipated from the increased drive, the reduction was achieved at the cost of a 5-db increase in conversion loss to the desired signal. This can be tolerated when adequate preamplification precedes the mixer.

D. INTERMODULATION AND CROSS MODULATION

   1. Source of Spurious Response

      a. Intermodulation

If two or more strong, unwanted signals enter the antenna of a receiver, spurious responses might occur due to the nonlinearity of the RF stages or the mixer. Odd-order products of the incoming signal cause the most trouble, since the even order products can be balanced out by use of a balanced mixer. The third order product causes most of the spurious responses since the amplitude generally goes down as the order increases.

If: $e_{in} = E_1 \sin W_1 t + E_2$ (III-9)

and the equation for the nonlinear stage is

$$e_o = a_o + a_1 e_{in} + a_2 (e_{in})^2 + a_3 (e_{in})^3 \qquad\qquad (\text{III-10})$$

A/B	ORDER														
	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
0	B	B±A	B±2A	B±3A	B±4A	B±5A	B±6A	B±7A	B±8A	B±9A	B±10A	B±11A	B±12A	B±13A	B±14A
1/16															15A
1/15														14A	
1/14													13A		
1/13												12A			
1/12											11A				2B-13A
1/11										10A				2B-12A	
1/10									9A					2B-11A	
1/9								8A					2B-10A		
1/8							7A					2B-9A			
2/15															14A-B
1/7						6A				2B-8A				13A-B	
2/13													12A-B		
1/6					5A				2B-7A			11A-B			
2/11											10A-B				2B-12A
1/5				4A				2B-6A		9A-B				3B-11A	
3/14															13A-2B
2/9									8A-B				3B-10A		
3/13														12A-2B	
1/4			3A				2B-5A	7A-B					3B-9A	11A-2B	
3/11												10A-2B			
2/7							6A-B				3B-8A				
3/10											7A-2B				4B-11A
4/13															12A-3B
1/3		2A				2B-3A / 3A-B				3B-7A / 4A-2B				4B-9A / 2B-3B	
4/11												10A-3B			
3/8									7A-B				4B-9A		
2/5				4A-B					3B-6A			9A-3B			
5/12															11A-4B
3/7								6A-2B				4B-8A			
4/9											8A-3B				5B-10A
5/11														10A-4B	
1/2	A			3A-B	2B-3A		5A-2B	3B-5A		7A-3B	4B-7A		9A-4B	5B-8A	
6/11															11A-5B
5/9											8A-4B				
4/7									6A-3B				5B-8A		
3/5					4A-2B					4B-6A				9A-5B	
5/8											7A-4B				6B-9A
2/3			2A-B					3B-4A	5A-3B			5B-7A	8A-5B		
7/10															9A-6B
5/7										6A-4B				6B-8A	
3/4					3A-2B					4B-5A		7A-5B			
7/9														8A-6B	
4/5							4A-3B				5B-6A				
5/6										5A-4B			6B-7B		
6/7											6A-5B				
7/8															7B-8A
1					2B-2A		3B-3A		4B-4A		5B-5A				

Figure III-9. Spurious Response Char
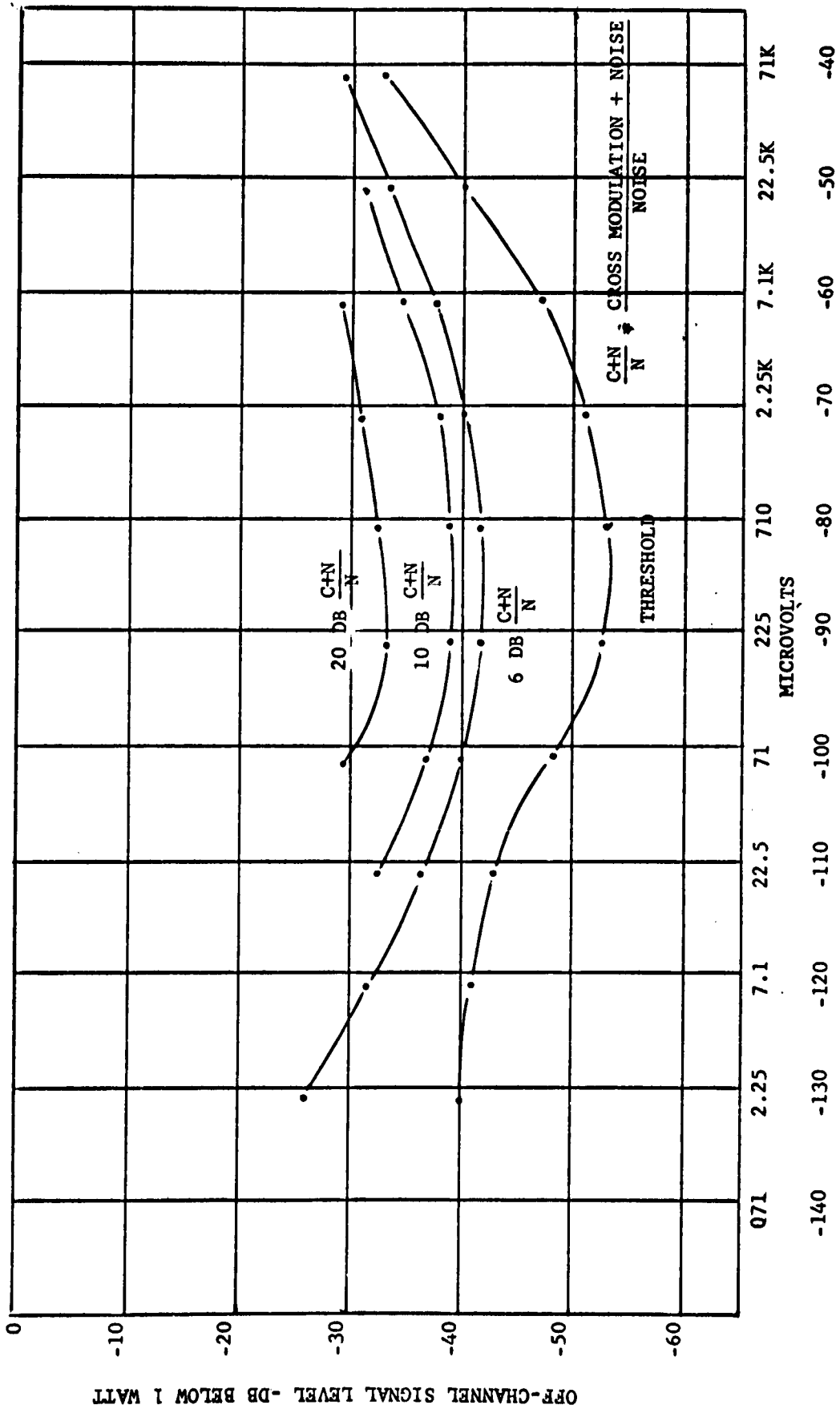
Figure III-12. Cross Modulation Characteristics of a Typical UHF Receiver

then the third order term is

$$e_3 = a_3 (E_1 \sin w_1 t + E_2 \sin W_2 t + E_3 \sin W_3 t)^3 \qquad \text{(III-11)}$$

which when expanded is:

$$e_3 = a_3 \left[\frac{E_1^3}{4} (3 \sin w_1 t - \sin 3 w_1 t) + \frac{E_2^3}{4} (\sin w_2 + -\sin 3 w_2 t) \right.$$

$$+ \frac{E_3^3}{4} (3 \sin w_3 t - \sin 3 w_3 t)$$

$$+ \frac{3 E_1^2 E_2}{2} (\sin w_2 t + 1/2 \left[\sin(2w_1 - w_2)t - \sin(2w_1 + w_2)t \right])$$

$$+ \frac{3 F_1^2 E_3}{2} (\sin w_3 t + 1/2 \left[\sin(2w_1 - w_3)t - \sin(2w_1 + w_3)t \right])$$

$$+ \frac{3 E_2^2 E_1}{2} (\sin w_1 t + 1/2 \left[\sin(2w_2 - w_1)t - \sin(2w_2 + w_1)t \right])$$

$$+ \frac{3 E_2^2 E_3}{2} (\sin w_3 t + 1/2 \left[\sin(2w_2 - w_3)t - \sin(2w_2 + w_3)t \right])$$

$$+ \frac{3 E_3^2 E_1}{2} (\sin w_1 t + 1/2 \left[\sin(2w_3 - w_1)t - \sin(2w_3 + w_1)t \right])$$

$$+ \frac{3 E_3^2 E_2}{2} (\sin w_2 t + 1/2 \left[\sin(2w_3 - w_2)t - \sin(2w_3 + w_2)t \right])$$

$$+ \frac{3 E_1 E_2 E_3}{2} (\sin(w_1 + w_2 - w_3)t + \sin(w_1 - w_2 + w_3)t - \sin(w_1 - w_2 - w_3)t -$$

$$\left. \sin(w_1 + w_2 + w_3)t) \right] \qquad \text{(III-12)}$$

As can be seen, there are a great many products produced. The difference terms such as $\sin(2w_1 - w_2)t$, $\sin(2w_2 - w_3)t$, $\sin(2w_1 - w_3)t$, $\sin(2w_3 - w_2)t$, $\sin(2w_2 - w_1)t$, $\sin(2w_3 - w_1)t$ as well as the fundamentals such as $\sin w_1 t$, $\sin w_2 t$ and $\sin w_3 t$ cause the spurious responses, since the sum terms are usually well out of the passband of the IF amplifier. It should be noted that these difference terms

can cause spurious responses even though the fundamentals that cause
them are not within the passband.

   b. Cross Modulation

   Cross Modulation is similar to intermodulation because they are
both caused by the nonlinearity of the front-end of the receiver. The
effect that it has on a receiver is that a wanted signal is modulated
by one or more unwanted pulse- or amplitude-modulated signals. There-
fore, the wanted signal has to be present for the effect of cross
modulation to be seen. It is usually encountered when several equip-
ments are operated within close proximity of each other.

   Figure III-12 gives typical cross modulation characteristics of
a UHF receiver. There is a 5-Mc separation between the wanted and
unwanted signals. The curves show the variation of receiver sensitivity
as a function of signal levels.

   2. Elimination of Spurious Response

   Spurious responses due to intermodulation and cross modulation can
be eliminated by the same techniques.

   a. Minimize Circuit Nonlinearities

   If an RF stage is utilized in the receiver front-end, care should
be taken to achieve a linear amplification characteristic. Also, the
mixer parameters should be picked in order to decrease the third-order
curvature. It can be shown that there is a definite relationship
between the amplitude of a given order difference frequency and the
same-order harmonic produced in a mixer, e.g.,

$$2f_{LO} \approx f_s - f_{LO} \qquad\qquad\qquad\qquad (III-13)$$

$$3f_{LO} \approx 2f_{LO} - f_s \qquad\qquad \text{Equations of amplitude only.} \quad (III-14)$$

$$4f_{LO} \approx 3f_{LO} - f_s \qquad\qquad\qquad\qquad (III-15)$$

   It is therefore possible to optimize the parameters for an RF
stage or a mixer with the use of only the local-oscillator frequency.
By varying the parameters of the RF stage, the harmonic generation
can be reduced to a minimum - especially the third order. The RF
stage will react in the same manner when there are two or more signals
present.

   b. Preselection.

   Filtering or preselection can be utilized between the antenna
and the first stage in order to reduce inter- and cross-modulation.
As mentioned elsewhere, the use of both filtering and a tunable
preselector is recommended. The filtering will attenuate a signal
outside the bandpass to a level approaching noise. The preselector,
which has inherent frequency-selective limiting, will produce a maximum

attenuation of 40 to 50 db within the overall band.

c. Gating

Gating could be utilized to eliminate inter- and cross-modulation but it is not a desirable means of rejection, since the entire receiver must be blanked.

E. INTERMEDIATE FREQUENCY WITHIN THE SIGNAL- OR THE LOCAL-OSCILLATOR FREQUENCY BAND.

1. Source of Spurious Responses

Other, more serious responses can arise for which preselection is ineffective. If the intermediate frequency is within the signal band, the entire receiver can be jammed by this frequency. Or, if the IF frequency is chosen so that the signal frequency is a low-order harmonic of the IF frequency, several spurious responses occur which cannot be eliminated by preselection. Figure III-13 demonstrates these spurious responses. Here, tne signal frequency range is 500-1100 kc and the IF is 455 kc. Spurious responses are seen to occur at 910 kc and the 1365 kc which cannot be eliminated by preselection. This is shown by drawing parallel lines on each side of 11(1).

Another possible response for which preselection is ineffective, is that which is produced by a local-oscillator harmonic which falls within the passband of the desired output.

2. Elimination of Spurious Responses

The only effective method for the elimination of the spurious responses described above is to properly choose the IF frequency. A discussion of the choice of IF frequency is found in paragraph C.2.b. of this appendix.

F. OTHER FACTORS AFFECTING RECEIVER FIDELITY

Two other factors which must be considered in regard to receiver fidelity and sensitivity are: (1) a reduction in sensitivity because of strong signals on or near the desired frequency; and (2) local-oscillator radiation, which creates a possible spurious response to other receivers.

1. Strong Signals on or Near the Desired Frequency

The chief disturbance caused by strong signals within the band is that the receiver will become desensitized and eventually saturated as the amplitude of the signal increases. Desensitization is defined as the reduction in gain for the wanted signal caused by the over-loading of a portion of the receiver by a strong unwanted signal. The amount of desensitization will depend upon the linearity of the re-ceiver. A good receiver will have little reduction in sensitivity until a point is reached where the receiver noise is no longer seen on the dis-play. At this point, the receiver sensitivity should decrease at the rate

of 1 db for every db of interference power injected. An example of typical results on a radar receiver is shown in Figure III-14.

One remedy for this form of interference is to use a high pass filter between the detector and the video amplifier to eliminate any DC component due to the interfering signal. This will prevent the video amplifier from overloading. Also, logarithmic or linear-logarithmic amplifiers can be used in place of the standard IF and video amplifiers, since a dynamic range of 60 db and greater can be achieved. The Lin-Log is preferable, in general, since the signal is discernable above noise more rapidly.

As mentioned previously, preselectors were recommended for image signal rejection. If the preselector has a limiting characteristic, such as is found in Yttrium-Iron-Garnet devices, these large signals can be attenuated.

Another means of strong signal rejection is to use a tunable band-reject filter placed between the antennas and the first RF stage. If the unwanted signal is at a known frequency, the operator can tune the band-reject filter to this frequency and thereby attenuate the unwanted signal. Approximately 60 db of attenuation is obtainable by this means, while causing less than 0.5 db insertion loss at the desired frequency.

Gating can be used as a last alternative to suppress strong signals but this method is not recommended since the display time is reduced. This is because the IF amplifier will have to be cut off an increment of time before the large signal reaches it and then there is a delay after the gate is removed for the amplifier to recover. A block diagram of the circuit that can be used for gating is shown in Figure III-15. A sample of the IF is taken from the adder, amplified, and then detected. If this detected IF is greater than a preset threshold level, bias is applied to the first two stages of the main IF amp or video amp. This bias reduces the gain of the amplifier as the strong signal increases and cuts the amplifier off if the signal becomes large enough.

2. Local Oscillator Radiation

Since the local oscillator generates a relatively high amplitude signal, radiation of the fundamental and its harmonics might be a source of interference to nearby receivers. One method of eliminating local oscillator radiation is to use an isolator between the antenna and mixer. This device has a low insertion loss in the forward direction and high insertion loss in the reverse direction. Also some suppression of the local oscillator frequency will be received due to the selectivity of the preselector. If RF amplifier stages are utilized, the local oscillator will be attenuated by the back to front isolation of these stages, in addition to the selectivity of the tuned circuits. A typical rejection of 20 db to 30 db per stage can be achieved in this manner.

The use of a tunnel diode as a converter with a tunnel diode supplying the local oscillator power diminishes the possibility of radiation, since the local oscillator power necessary for efficient converter operation is approximately 20 db less than that required by a crystal mixer.

The standard balanced mixer circuit affords approximately 20 db of local oscillator path to signal path isolation. Thus, local oscillator radiation is reduced in a balanced mixer circuit.

APPENDIX IV

CHOICE OF IF FREQUENCY

It has been shown (Appendix I) that the choice of IF frequency is extremely important in the minimization of spurious responses. However, other factors, besides spurious response rejection, which must be considered in the choice of IF frequency are:

Noise Figure

Criticalness of Circuit

Separation of Video from the IF

Information Bandwidth Requirements

Noise and Spurious Responses Due to the Local Oscillator

A. NOISE FIGURE

For a given amplifier, lower noise figures can be achieved by using a lower IF. This is due to the fact that the optimum source resistance decreases, therefore the input circuit bandwidth increases as the frequency is increased, thus causing higher noise figures. If, on the other hand, the bandwidth is made greater than that produced by the optimum source resistance, the relative noise figure advantage of a low intermediate frequency is not as great as it is with narrower bandwidths.

B. CRITICALNESS OF CIRCUIT

For lower intermediate frequencies, the reactive variations due to changing the active portions of a circuit have less effect for a given bandwidth. The added capacity used for tuning can be greater, therefore the capacity added by the active element such as a tube or transistor has less effect on the overall circuit. The probability of having to retune or rework a circuit when the active element is changed is therefore lessened when a low IF is used.

C. SEPARATION OF VIDEO FREQUENCIES FROM THE INTERMEDIATE FREQUENCIES

For a given video and IF bandwidth, a high intermediate frequency reduces the possibility that the IF components of the rectified signal may get into the video amplifier. With care in decoupling, this need not be a problem.

D. BANDWIDTH REQUIREMENTS

If a large bandwidth is needed to produce an accurate reproduction of the waveform of an incoming signal, the use of a high intermediate frequency may be necessary if the design of each stage is to be possible (the bandwidth of each stage of an amplifier using cascaded tuned circuits must be greater than the overall bandwidth).

E. NOISE AND SPURIOUS RESPONSES OF THE LOCAL OSCILLATOR

In the case of wide-range tuned oscillators, such as the backward-wave oscillator, spurious outputs other than the center frequency might be generated. These spurious oscillations near the main local-oscillator frequency might be caused by oscillation in a nearby mode with subsequent intermodulation with the main signal, or by low-frequency oscillation within the tube. The latter may cause frequency modulation of the main signal. In either case, the pattern will appear as a series of discrete, evenly spaced signals which resembles the characteristic of an FM spectrum. Along with these spurious oscillations, the local oscillator generates noise, centered on the main frequency, which decreases in amplitude as its frequency increases or decreases from the main frequency. If the IF is made large, the possibility of these responses getting within the RF passband is reduced, since noise and spurious-response levels increase as the local-oscillator frequency is approached.

F. SUMMARY

As can be seen, the noise figure and the criticalness of the circuit call for a low IF, while spurious-response rejection, separation of the video from the IF, large information bandwidth, and local-oscillator noise all require a high IF frequency. In consideration of these factors, a compromise will be made in order to achieve an optimum IF. This compromise is discussed in Appendix V.

APPENDIX V

CHOICE OF RECEIVER FRONT-END

A. GENERAL

The choice of intermediate frequency (and, therefore, local-oscillator frequency) has been discussed previously in Appendix IV, and was shown to effect spurious responses, noise figure, criticalness of circuit, separation of video from IF, information bandwidth available, and noise and spurious responses due to the local oscillator.

From a spurious-response consideration alone, the values of local-oscillator frequency and intermediate frequency presented in Table III of the text can be utilized. The frequency values of Table III were calculated on the assumption that no spurious responses of less than the 10th order existed. Calculations were made for both single-ended mixers and balanced mixers (where approximately 50 db attenuation of certain even-order harmonics is achieved).

Figure 1 of the text, shows the values of local-oscillator and intermediate frequencies which will not produce spurious responses of less than the 10th order when single-ended mixers are used.

B. COMPARISON OF APPROACHES

1. Case I

Figure 1 shows the frequency range of each band division. The bands mentioned in the following discussion will refer to the bands indicated in this figure. In Case I, (Figure 1) the entire local-oscillator frequency range is above the signal frequency range. This allows excellent image-frequency rejection and local-oscillator radiation suppression, since the input band-pass filter will greatly attenuate frequencies so far removed from the passband. Moreover, if the input filter were to have a harmonic response at the local-oscillator frequency, a band-reject filter could be utilized.

A tunable band-pass filter is required to prevent strong signals within the RF passband from saturating the preamplifier and to prevent spurious responses due to harmonics of signals within the RF passband and outside the IF passband.

In band 7 and above, double superheterodyne techniques must be employed since the allowable IF becomes too high for practical logarithmic, transistorized IF amplifiers.

Above band 15, the local oscillator can no longer be kept outside the signal-frequency passband because the swept local oscillator which would be required will not be realizable in the near future via solid state techniques.

The major disadvantages of having a local oscillator outside the passband are that (1) a high intermediate frequency necessarily results and (2) local oscillator-preselector tracking becomes difficult due to the large separation between signal and image frequency.

A high intermediate frequency is undesirable for two reasons. First, solid state implementation of a linear-logarithmic IF is extremely difficult at high frequencies. Second, the narrow IF bandwidth required for sector scanning is difficult to realize in a high frequency IF due to the high loaded Q required.

While the local oscillator is changing frequency 1.05:1, the preselector is changing frequency 2:1. This makes local oscillator-preselector tracking difficult. This tuning-sensitivity difference can, however, be overcome through the use of equalization circuitry.

Another technique which has been investigated to minimize the tracking problem is the utilization of a swept local oscillator which is made by beating a high-frequency fixed oscillator with a swept, low-frequency oscillator. A possible local oscillator for band 6 is shown in Figure V-1.

The swept local-oscillator frequency must be in the 10- to 14- Mc range in order to provide adequate filtering from the signal band. A swept local-oscillator frequency which ranges below 4 Mc cannot be utilized, since a 4-Mc swept range is then not available without entering the signal band.

Since 10 to 14 Mc is closer to 4 to 8 Mc, than 84 to 88 Mc, the tracking problem has been minimized. However, spurious responses can now be generated as shown in Figure III-5 of Appendix III, for when the swept local-oscillator frequency equals 12.33 Mc, A/B = 1/6 and the 7th-order spurious responses will be developed. For example, assume that the signal frequencies changes slightly to F sig = 12 Mc. Then B+A = 86 Mc for the generated local-oscillator frequency. However, 2B-5A and 7A are also generated, producing:

$$2B - 5A = 88 \text{ Mc}$$

$$7A = 84 \text{ Mc}$$

Since the local-oscillator power required at 86 Mc is relatively high (approximately 1 milliwatt), the drive power at 74 Mc and 12 Mc must also be high. Therefore, as shown in Figure III-8, the 7th-order response may be only on the order of -30 db to -40 db below the desired response.

2. Case II

The local-oscillator frequency band has been allowed to extend down into the signal frequency band in Case II. While the local oscillator-preselector tracking problem is now minimized, local-oscillator radiation and image-frequency rejection become a problem.

Local-oscillator radiation can be reduced to less than -67 dbm (the level required by MIL-I-26600) via active RF amplifiers with an inherently high reverse isolation. In addition, the preselector required for image rejection will add to the local-oscillator suppression.
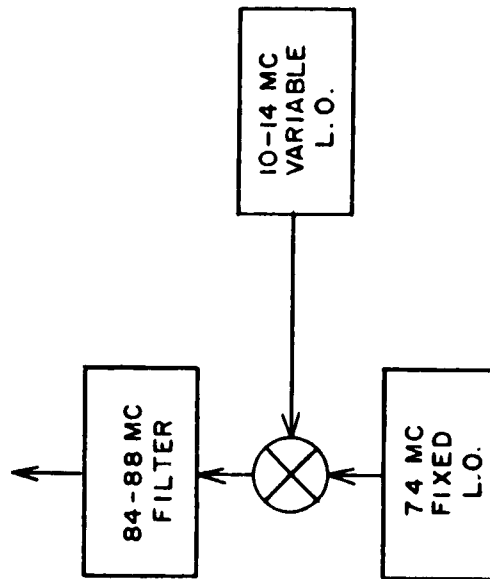
Figure V-1. Typical Swept Local Oscillator to Minimize
Tracking Problem When High Frequency if is Utilized

A preselector filter must be used for image-frequency rejection.
Since a tunable filter is required to eliminate spurious responses from
signals which are within the RF passband, but outside the IF passband,
the only stress which image-frequency rejection places on the preselector
is the requirement for additional filter sections to steepen the skirt
selectivity. It must be noted that the addition of filter sections may
prove difficult due to the increased problem of tracking the additional
filter sections to each other and to the local oscillator.

The single-superheterodyne approach is utilized throughout Case II.
Above 500 Mc, a 95-Mc IF is used in each band, since 95 Mc represents
the highest IF presently attainable in a transistorized lin-log ampli-
fier which is capable of providing the necessary bandwidths.

3. Case III

Above 500 Mc, the value of the allowable local-oscillator frequen-
cies in Case II, as shown in Figure 1 of the test, continues to increase
above 95 Mc. Therefore, a double-superheterodyne front-end is utilized.

The circuits shown in Case III require an RF preamplifier to reduce
the noise figure to a level that is comparable with that of a single-
superheterodyne receiver. The first-IF frequencies in Case III are the
maximum allowable. If a lower first-IF frequency is utilized, (on the
order of 200 Mc) an RF preamplifier may not be necessary because a low-
noise transistorized IF amplifier can be used.

C. CHOICE OF APPROACH

Reference to Table I of the text, shows that the basic receiver
approach which offers the optimum compromise and the most versatility
is the single-superheterodyne receiver (with double superheterodyne
used above 500 Mc). Image response and antenna radiation can be sup-
pressed to an adequate level by using devices such as tunable pre-
selector filters and image rejection mixers, and by choosing a high IF
frequency. Signal intercept probability, sensitivity, and selectivity
can all be maximized through utilization of a variable sweep rate and
variable IF bandwidth. Moreover, cost, size, complexity, and power
consumption can be minimized by the use of solid state oscillators.

The single-superheterodyne receiver of Table V-1, Case I, was
shown to have the following advantages and disadvantages:

Table V-1. Case I (LO Band Above Signal Band)

ADVANTAGES	DISADVANTAGES
1. Eliminates image problem.	1. High local-oscillator frequency.
	2. Preselector-local oscillator tracking problem.
	3. High IF frequency; difficult to obtain resolution.
	4. RF preamplification is required above 8 Mc in order to reduce the double-conversion noise figure.
	5. Not presently practical to implement above 4 Gc.

Table V-2, Case II, where the local-oscillator frequency range was allowed to extend into the signal band, was shown to have the advantages and disadvantages listed below:

Table V-2. Case II (LO Band Inside Signal Band)

ADVANTAGES	DISADVANTAGES
1. Minimize preselector - local oscillator tracking problem. 2. Single approach throughout the frequency band. 3. Lowest possible local oscillator and intermediate frequencies utilized. 4. No preamplification required to produce a sensitivity of -90 dbm.	1. High loaded Q required in the tunable preselector in order to eliminate image response

It therefore appears clear that if adequate preselection can be provided, the single-superheterodyne approach outlined in Table V-2, Case II, provides the optimum approach.

Table V-2, Case II, as it stands, however, presents one problem when applied to a practical implementation of the receiver; namely, the multiplicity of intermediate frequencies. Unless the number of intermediate frequencies is reduced, implementation of AFC circuitry and automated circuitry in each band would become unreasonable due to the many different frequency designs required.

Actually, the number of intermediate frequencies required can be reduced to five, as shown in Figure 2, of the text.

In Band 1, the 18-kc IF will be converted to the IF of Band 2 (45 kc). This will make the fabrication of the IF amplifier more practical, as well as reduce the total number of intermediate frequencies. There will be no increase in noise figure since an RF stage will be required for preselection in any case.

Since Band 3 has a maximum IF of 95 kc, 45 kc can be utilized here also, if adequate image rejection can be obtained. Assuming four sections of preselection, the image rejection can be calculated from:

$$B = \frac{2 \Delta f}{4 \tau}$$

(V-1)   Where B = preselector bandwidth

Δf = 2 x intermediate frequency (difference between center frequency and the frequency at which the suppression is desired.

τ = rejection expressed as a voltage ratio.

For an IF of 45 kc,

$$15 = \frac{180}{4\sqrt{\tau}} \qquad \text{(V-2)}$$

$$4\sqrt{\tau} = 12$$

$$\tau = 20,736 \approx 86 \text{ db}$$

Thus, 45 kc can be utilized as an IF for Band 3. If a 45-kc IF becomes impractical, Figure 3 shows an alternate approach for Bands 1 to 3. However, all the disadvantages of Table V-2, Case II, as outlined previously, accompany this alternate approach.

A similar analysis for Band 4 shows that 45 kc is too low for an IF in this band. Therefore, 175 kc is utilized.

A 175-kc IF is also adequate for Band 5 as shown by:

$$50 = \frac{700}{4\sqrt{\tau}} \qquad \text{(V-3)}$$

$$4\sqrt{\tau} = 14$$

$$\tau = 38,400 \approx 91 \text{ db}$$

Analysis of Band 6 shows that a 175-Mc IF cannot be utilized. Therefore, a 700-kc first-IF is utilized and the 700-kc signal is down-converted to 175 Mc. Again, RF stages are necessary for preselection and therefore no noise-figure degradation results.

A similar study was conducted for Bands 7 through 12. A total of four intermediate frequencies is therefore required for the first 12 bands. Note: Double conversion to a single IF cannot be utilized for all twelve bands, since inadequate frequency resolution would result.

Bands 13 through 20 employ a 95-Mc IF. This is possible since the preselector bandwidth requirements change only from 15 Mc to 30 Mc. For 30 Mc,

$$30 = \frac{380}{4\sqrt{\tau}}$$

$$4\sqrt{\tau} = 12.6$$

$$\tau = 25,600 \approx 88 \text{ db}$$

If a 120-kc bandwidth proves impossible to implement at 95 Mc, a second conversion to a 30- to 60-Mc IF will be utilized. However, a second conversion may require the addition of an RF preamplifier.

NOISE FIGURE OF A DOUBLE-SUPERHETERODYNE RECEIVER

With reference to Figure V-I-1, the double-conversion type receiver can be considered as a cascade circuit which consists of two networks: Network No. 1, consisting of the input band-pass filter and preselector followed by the first mixer, and Network No. 2, consisting of the converter output filter, the second mixer and the IF amplifier. The total noise figure of these networks can be computed from:

$$F_T = F_1 + \frac{F_2 - 1}{G_1} \qquad \text{(V-I-1)}$$

where F_1 = Noise figure of Network No. 1

F_2 = Noise figure of Network No. 2

G_1 = Available gain of Network No. 1

Assuming negligible losses in filter No. 1, the noise figure of network No. 1 and No. 2 is given by:

$$F_1 = L_{c/tm} \qquad \text{(V-I-2)}$$

and $$F_2 = L_i L_{c2}(t_m + F_{IF} - 1) \qquad \text{(V-I-3)}$$

respectively, where:

L_c = Conversion loss of the mixer*

t_m = Equivalent noise temperature of the mixer

L_i = Insertion loss of filter No. 2

F_{IF} = IF noise figure

Now $1/G_1$ in equation (1) equals L_c.

Thus: $$F_T = L_c [t_m - 1 + L_i L_c (t_m + F_{IF} - 1)] \qquad \text{(V-I-4)}$$

Typical figures for L_c, t_m, t_i and F_{IF} are:

L_c = 5 (7 db) for mixers between 1 Gc and 10 Gc

t_m = 1.3

L_i = 1.3 (1 db)

F_{IF} = 2.5 (4 db) for an IF bandwidth of approximately 20 Mc.

---

* Assume both mixers have identical conversion loss and noise temperature. The expression for F_T can readily be modified, if the mixers have different characteristics.
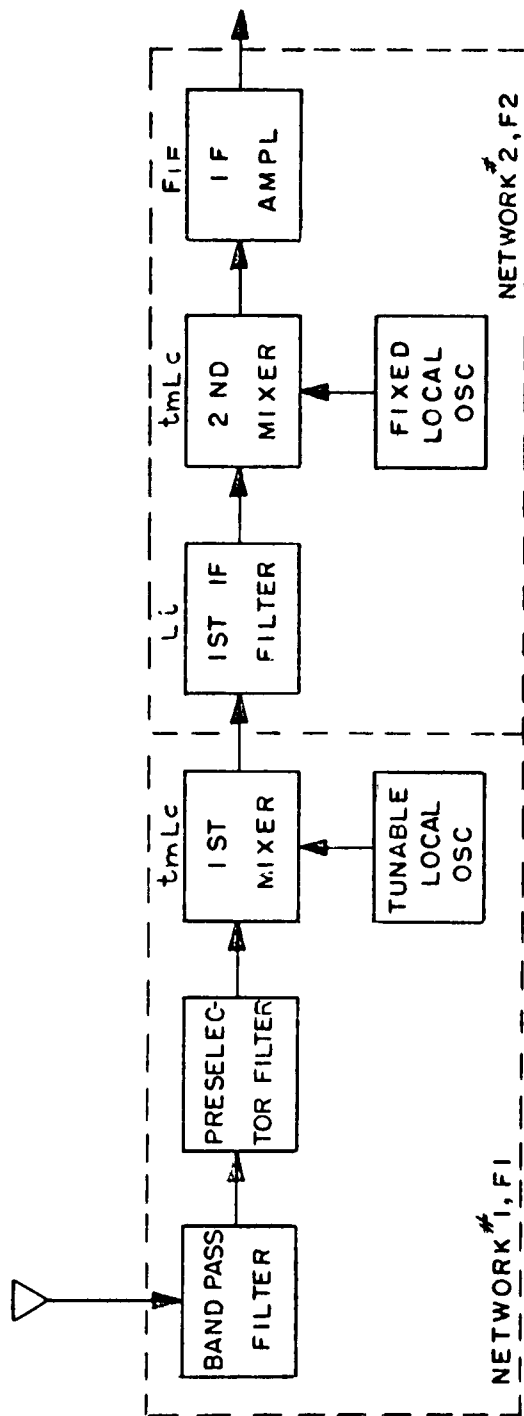
Figure V-1-1. Typical Double Superheterodyne Receiver.

1101458

Inserting all the above values into (V-I-4):

$$F_T = 5[1.3 - 1 + (1.3)\ (5)\ (1.3 + 2.5 - 1)]$$

$$F_T = 92.5 = 19.7 \text{ db} \approx 20 \text{ db*}$$

The above example clearly shows that F_T can be approximated by:

$$F_T \approx L_i L_c^2 F_{IF} \qquad\qquad (V-I-5)$$

The noise figure computed from Equation (5), using the above values of L_i, L_c and F_{IF}, is:

$$F_T \approx (1.3)\ (5)^2\ (2.5) = 81.25 = 19.1 \text{ db}$$

which is close to the correct value obtained from Equation (V-I-4). Equation (V-I-5) shows that the minimum discernable signal power is approximately proportional to the insertion loss of the converter output filter and the IF noise figure.

For a single-superheterodyne receiver, again assuming negligible insertion loss due to the input filters, the total receiver noise figure is given by:

$$F_T = L_c\ (t_m - F_{IF} - 1) \qquad\qquad (V-I-6)$$

Using the values of L_c, t_m, and F_{IF} given above:

$$F_T = 5(1.3 + 2.5-1) = 14 = 11.5 \text{ db}.$$

Thus, the single-superheterodyne approach affords approximately a 9.0-db noise figure advantage over the double-superheterodyne receiver.

With preamplification, however, the noise figure difference between the single- and double-conversion receivers may be made insignificant. For instance, assume a preamplifier having an 8-db noise figure and 20 db of gain (values which are readily obtainable from traveling wave tube and tunnel diode broadband amplifiers). Then from Equation (V-I-1) where the symbols now represent:

$$F_1 = \text{noise figure of preamplifier}$$

$$F_2 = \text{total receiver noise figure}$$

$$G_1 = \text{gain of preamplifier}$$

The total noise figure for the single- and double-superheterodyne receivers are:

$$F \text{ single} = 6.3 + \frac{14-1}{100} = 6.44 = 8.1 \text{ db}$$

and

$$F \text{ double} = 6.3 + \frac{92.5-1}{100} = 7.23 = 8.6 \text{ db, respectively.}$$

---

* If the input filter losses are not negligible, these losses must be added to the total noise figure derived.

Therefore, with 20 db of preamplification gain, the noise-figure disadvantage of a double-superheterodyne receiver is reduced to less than 0.5 db. With greater than 20 db of preamplifier gain, the noise-figure difference is reduced still further.

The conclusion is therefore reached, that if preamplification is necessary with a single-superheterodyne receiver, no appreciable noise-figure degradation will be introduced by replacing the single-superheterodyne receiver with a double-superheterodyne receiver.

APPENDIX VI

INTERCEPT PROBABILITY

A. GENERAL

The approach used in this investigation of Intercept Probability, Acquisition Probability and Probable Acquisition Interval is one of a relatively simple problem in coincidence of two (or more) simultaneous events. A mechanical analog is used to show that the analysis is not limited to radio receivers nor to a particular type or model. Some basic requirements are deduced from the mechanical analog and are related to the problem of the panoramic (time-shared) receiver receiving uniform pulse signals from a scanning antenna. Although no general equation for Acquisition interval can be derived, several assymptotic equations are developed.

A Mathematical Analysis of the Probability of Acquiring a signal on a single random sweep of the receiver is presented in Appendix VI-I. A discussion and analysis of the Probable Acquisition Interval is presented in Appendix VI-II.

B. MECHANICAL ANALOGY OF THE PROBLEM

Terms Used

α (Alpha) — beam angle of transmitting antenna above reference-power level which produces a threshold signal at the receiver.

B_r — total range of center frequency covered in one scan of the receiver.

b_r — acceptance bandwidth of the receiver at an assumed threshold level of sensitivity

b_s — bandwidth of the components of the signal

D_a — the duty cycle of the antenna

D_s — the study cycle of the signal (t_s/T_s)

f_H — frequency of acquisition of the signal

f_o — carrier frequency of the incoming signal under study

f_r — frequency of receiver scan $(1/T_r)$

f_m — maximum receiver sweep frequency that enables the acquisition of one pulse of a given signal on every sweep.

f_{rma} — maximum receiver sweep frequency that enables the acquisition of at least one pulse of a given signal from a scanning antenna on every sweep of the receiver.

f_s — pulse repetition frequency of signal $(1/T_s)$

f_{sa} — maximum receiver sweep frequency that will enable one acquisition per receiver scan for signals from a rotating antenna

F_1 - lowest frequency tuned by the receiver

F_2 - highest frequency tuned by the receiver

k   - the ratio of active receiver sweep time to sweep period.

N_{100} - number of sweeps required for 100% certainty of seeing one pulse from any single radiator

$p^{(1)}$ - the probability of seeing one coincidence of events

T_H - time between two successive acquisitions of the signal

θ (Theta) - scan angle of the transmitting antenna

T_r - time duration from start of one receiver sweep to start of next sweep period of the transmitted signal

t_r - time duration in each receiver sweep that receiver can receive a particular signal frequency

T_{rm} - minimum receiver sweep period for acquisition of one pulse of a given signal on every sweep ($T_{rm} = 1/f_{rm}$)

t_{rs} - total time during each sweep of the receiver that a pulsed signal at a particular frequency can be acquired (t_{rs} t_r)

T_s - period of the transmitted signal

t_s - duration of signal pulse

T_{ta} - time duration of each antenna scan

t_{ta} - time during each scan of the transmitting antenna that a signal can be detected by the receiver.

t_1 - t_{rs} - t_r

UCP   - Unit Coincidence Probability - the probability of acquiring one pulse on a single random sweep of the receiver.

The probability of acquisition of pulsed signals by a frequency swept receiver is explored in this report purely on the basis of coincidence of the pulse of RF energy arriving at the time the receiver tunes the proper input frequency. A mechanical analogy to the problem can be given to show that some parts of the problem are related to the characteristics of the receiver, and others are related to the characteristics of the signal and transmitting antenna.

Refer to figure VI-1 for a mechanical coincidence problem. The small turntable is the receiver which must indicate whether marbles are being dropped from the sorting machine on the upper turntable. When any part of a falling marble touches any part of the switch treadle on the receiving turntable, the lamp lights, indicating a "hit".

The mechanical coincidence can be directly related to the case of the panoramic (time-shared) receiver seeking a radar signal. A marble dropping mechanism, not available from AEL stock, drops marbles at a steady time interval, T_s, which is comparable to the pulse interval of
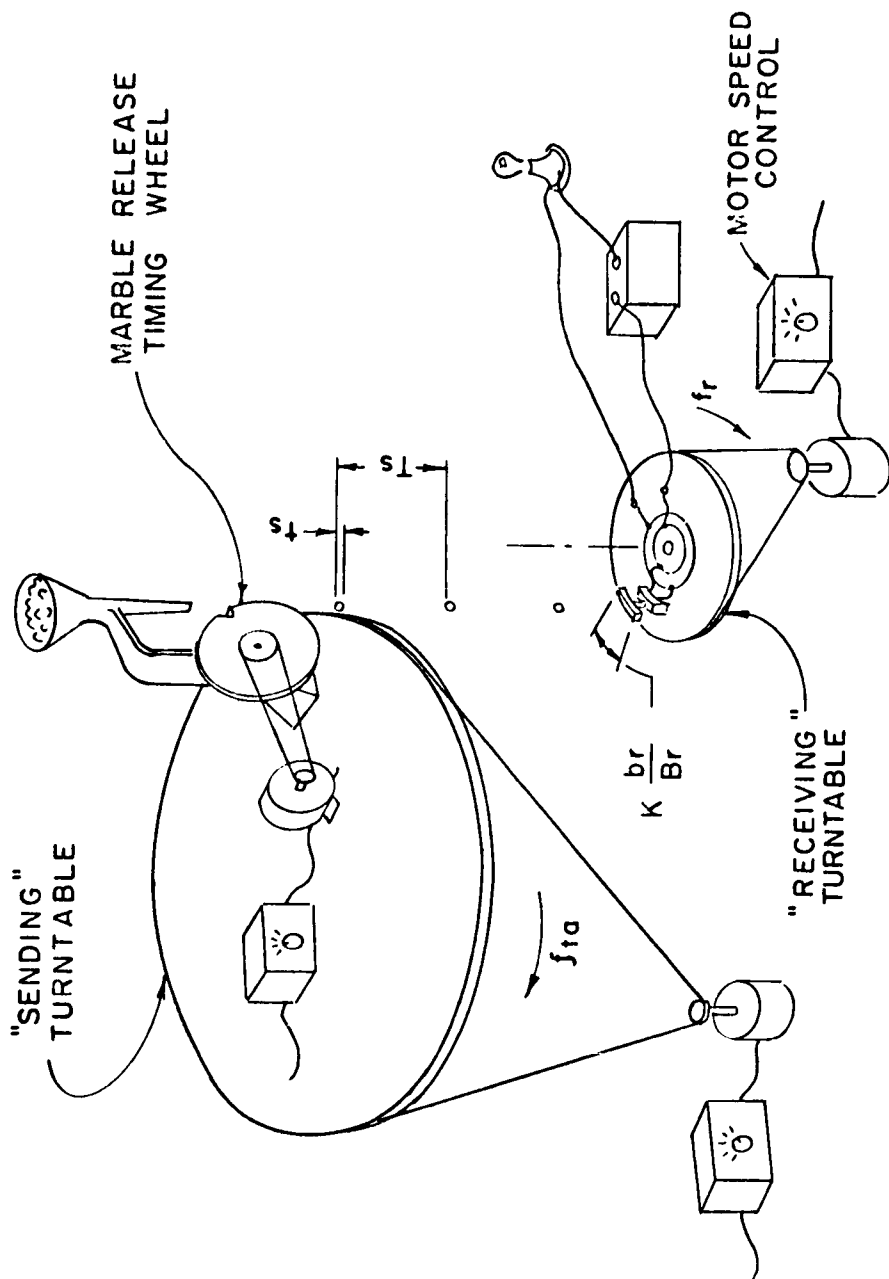
Figure VI-1. Mechanical Coincidence Problem

a radar set. The size of the marble may be compared to the time duration, t_s, and frequency spread, b_s, of the pulse itself. The rotation of the large turntable permits marbles to be dropped over the receiving mechanism for only a small part of its rotation, just as a radar antenna aims pulses toward a panoramic receiver over only a portion of the antenna sweep. The receiving turntable rotates. Once each revolution it places the switch treadle in a position where it could receive a marble. The arc of the switch is comparable to the acceptance bandwidth, b_r, of the panoramic receiver, and the circumference of the circle traversed by the switch is comparable to the total range of frequency, B_r, covered by a sweep of the panoramic receiver.

To start the problem, stop the rotation of both turntables in the positions shown in figure VI-1. Every marble dropped from the marble timer will hit dead center on the switch treadle and the lamp will flash on every marble. This condition is analogous to having a manually tuned receiver tuned to the "carrier" frequency of a pulsed signal from an antenna directed toward the receiver. The interval between light flashes is the same as the marble-to-marble time interval, T_s.

Next, the motor control of the lower (receiving) turntable is adjusted to turn very slowly. Now the only time the marbles can actuate the switch is during the time any part of the treadle is under some part of the falling marble.

Figure VI-2 shows the two limiting positions of the switch treadle that would be close enough to the falling marble to cause "hits". If the rotation of the receiving turntable is slow enough, the treadle will remain under the line of fall for a time which is greater than the time interval between drops, and at least one marble will hit the treadle on every rotation of the receiving turntable.

If it is assumed that the treadle width is b_r, and the width of the marble is b_s, then the distance traveled by the treadle while a successful hit is possible is:

$$X = \frac{b_r}{2} + b_s + \frac{b_r}{2} + b_r + b_s \qquad (VI-1)$$

If the radius of the turntable to the center line of the switch treadle is R, the circumference of the circle traversed by the treadle is $2\pi R$ making the total time of possible successful hits per revolution

$$t_x = \frac{X}{2\pi R} \ T_r = \frac{b_r + b_s}{2\pi R} \ T_r$$

if we let $2\pi R = B_r$ and $T_r = \frac{1}{f_r}$, then

$$T_x = \frac{b_r + b_s}{B_r f_r} \qquad (VI-2)$$

where T_r is the period of the turntable rotation.
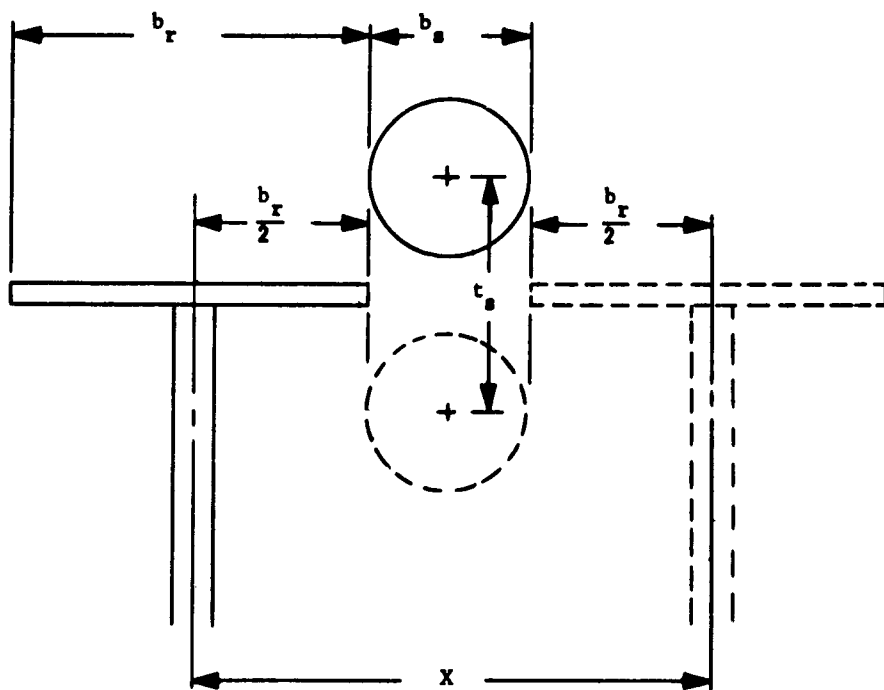
Figure VI-2. Limiting Position for Hits

If the time between marble drops is T_s, and the rotation speed of the turntable is f_r revolutions per unit time, the maximum turntable speed which would allow one hit per revolution would require $t_x \geq T_s$.

$$\frac{b_r + b_s}{B_r f_r} \geq T_s$$

$$\left[f_{rm} = \frac{b_r + b_s}{B_r T_s} \quad \frac{b_r + b_s}{B_r} \quad f_s \right] \qquad \text{(VI-3)}$$

In this condition, the rotational speed of the receiver is much slower than the rate at which marbles are being released. Many marbles will fall during the time the treadle is not under the line of drop. As the receiving turntable makes faster sweeps, there can be speeds at which the treadle always intercepts a marble. This will occur in every case where the receiving turntable makes a revolution in the exact time required for a given number of marbles to be released (e.g. one revolution per ten marbles dropped) or, when the receiving turntable makes one revolution in the time (t_s) required for each marble to fall a distance equal to its own diameter. Then, T_r t_s, the time required for the falling marble to pass the turntable.

In either case, a marble will be intercepted on every sweep of the receiving turntable.

The following condition we shall call synchronous.

$$T_s = n T_r = \frac{n}{f_r} \qquad \text{(VI-4)}$$

where n is an integer or the reciprocal of an integer.

This equation is significant when:

$$f_r > \frac{b_r + b_s}{B_r} f_s \qquad \left| f_s = \frac{1}{T_s} \right.$$

With exactly synchronous rotation speed of the receiving turntable, another undesirable but more probable condition may occur wherein the marble always misses the treadle. It is this condition which must be avoided, since in that case marbles could fall continuously without any indication from the receiver lamp.

Rotating the sending turntable introduces new effects. It will be assumed that the number of marbles falling in a given time remains the same at all speeds of rotation of the sending turntable, and that the marbles continue to fall straight down rather than being thrown outward by centrifugal force.

If the radius from the center of the sending turntable to the marble chute is P, and the width of the receiving treadle is X, the

central angle subtended by spread of possible successful falling marbles can be taken as:

$$\alpha \approx \sin^{-1} \frac{b_s}{P}.$$

let the total angle covered in a rotation of the sending turntable be $\theta = 2\pi P$.

The number of marbles dropped in one rotation of the sending turntable will be

$$N = \frac{T_{ta}}{T_s} \tag{VI-5}$$

the number of marbles dropped in the area of possible successes will be

$$N \frac{\alpha}{\theta} = \frac{\alpha}{\theta} \frac{T_{ta}}{T_s} \tag{VI-6}$$

and the time per revolution that marbles can fall in the area of possible success will be t_{ta} where

$$t_{ta} = \frac{\alpha}{\theta} T_{ta} \tag{VI-7}$$

For this discussion, it will be assumed that $T_s < t_{ta}$ with the result that one or more marbles will fall in the area of possible success on each revolution. In order to assure that at least one marble will strike the treadle on every revolution of the receiving turntable, the treadle must remain in the area of possible success for a period at least as long as one revolution of the sending turntable. The fastest allowable rotation of the receiving turntable for these conditions can be determined.

$$T_{ta} \leqq \frac{b_r + b_s}{B_r} T_r; \quad f_r \frac{1}{t_r}; \quad f_{ta} = \frac{1}{T_{ta}}$$

$$f_{rma} \leqq \frac{b_r + b_s}{B_r} f_{ta} \tag{VI-8}$$

The switch is sure to be struck by one or more marbles for every rotation of the receiving turntable. However, even though marbles are being dropped in the area of possible success, another hit cannot possibly be made for the time that the treadle is outside the area.

$$T_{NH} = \frac{B_r - b_r}{b_r} T_a \tag{VI-9}$$

For comparison, a typical panoramic receiver operating in the S-Band, sweeping a frequency range (B_r) of 2 G_c, with an acceptance bandwidth (b_r) of 20 Mc, receiving signals from an antenna scanned once a minute would require:

$$T_H = \frac{2000 \text{ Mc} - 20 \text{ Mc}}{20 \text{ Mc}} \cdot \frac{1}{1 \text{ rpm}}$$

$$= 99 \text{ minutes (see also Equation VI-I-18)}$$

Obviously, even though the probability of getting a hit on every sweep is 100%, this is not a desirable solution to the acquisition requirements since the receiver would scan only 5 times in one 8 hour day.

A second way which assures that an acquisition will be made for every marble dropped is to rotate the receiving turntable so fast that one full revolution bandwidth, B_r, in the time the shortest possible signal pulse will last. On many bands this would require sweep times of a few nanoseconds per sweep with flyback times becoming so formidable that triangular or even sinewave sweep would be required. The problem of loss of sensitivity due to effective filter bandwidth stretching and the problems involved in sweeping such a rate would probably make the wide-open crystal video receiver a more attractive choice. Study of this type of receiver will be discontinued at this point as being unreasonable for the present program.

C. PROBABLE ACQUISITION INTERVAL

The entire effort of this report is concerned with setting forth the requirements for the most frequent acquisitions. The preliminary investigation, based on an extensive study of the probability of intercept, has suggested that a sweep speed above the flicker rate, but also substantially below the PRF of typical signals for each band, would be adequate for most applications (refer to Appendix VI-II, figure VI-II-3). Curves have been plotted to show an Approximate Acquisition Interval, as shown in figure VI-3 which is based on the assumption: that to be 100% certain of an occurrence having probability

$$T_H \approx \frac{1}{p^{(1)}} \quad T_r \tag{VI-10}$$

The equation is invalid for uniform sweep speeds where the knowledge of the presence of an acquisition on any one sweep establishes the condition of probability for each successive sweep. By starting with a particular sweep on which an acquisition is obtained, it should be possible to derive exact or empirical expressions for subsequent acquisitions.

The assumption made in constructing figure VI-3 is that the number of random single sweeps which must be made to be 100 per cent certain of an acquisition is:

$$N_{100} = \frac{1}{p^{(1)}}$$
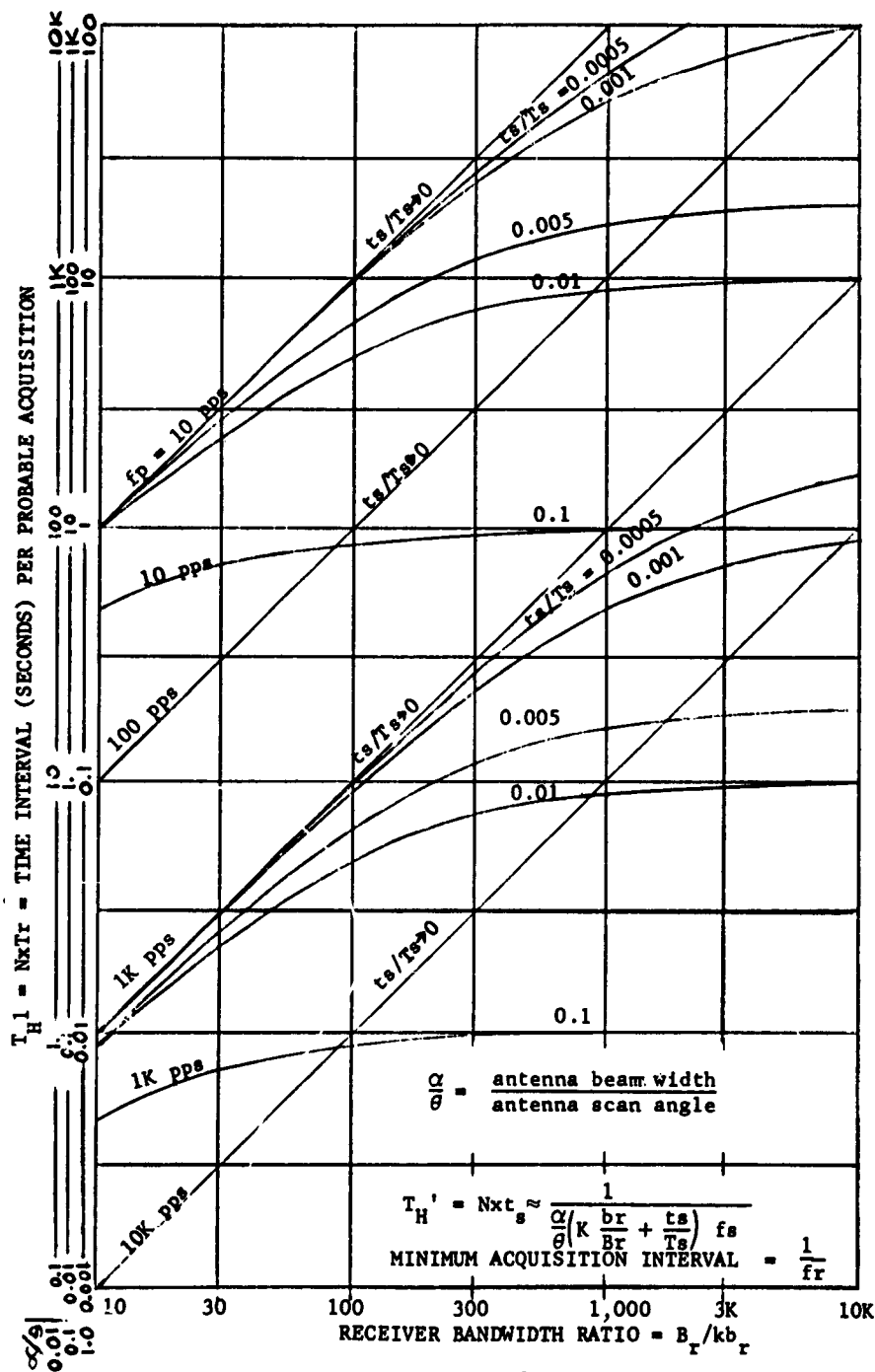
Figure VI-3. Approximate Acquisition Interval if Based on Probability for a Single Sweep. Pulsed Signals from Scanning Antenna Received by Frequency - Swept Receiver.

$$\frac{\alpha}{\theta} = \frac{\text{antenna beam width}}{\text{antenna scan angle}}$$

$$T_H' = Nxt_s \approx \frac{1}{\frac{\alpha}{\theta}\left(K\frac{br}{Br} + \frac{ts}{Ts}\right)fs}$$

MINIMUM ACQUISITION INTERVAL $= \frac{1}{fr}$

and that the total time to make N_{100} random sweeps would be

$$T_{H'} \geqq N_{100} \times T_r$$

A study of T_H based on beat-frequency concept shows that the value of T_H found in this way may be longer than the actual value.

The calculations on which figure VI-3 is based show generally that for a given receiver bandwidth ratio ($K\, b_r/B_r$) and a given signal PRF, there would be a nearly constant interval of acquisition which would be the same as the sweep interval for f_{rm}. For typical radar signals:

$$f_{rm} = k \; \frac{b_r}{B_r} \; f_s$$

and for faster sweeps, as stated above, it is approximately true that

$$T_{H'} = \frac{1}{p^{(1)}} \quad T_r = K \; \frac{b_r}{B_r} \; f_s \qquad\qquad \text{(VI-11)}$$

The most obvious disproof of the value of T_H' as the exact acquisition interval is shown in the case of a signal where $n f_r = f_s$, "n" being any positive integer. This is the condition called synchronous.

The repeat interval would be equal to the actual sweep time $(1/f_r = T_r)$ if the signal were acquired at all, and the probability of acquisition would be in accordance with equation (VI-I-11).

For nonsynchronous signals, the signal is always sure to be acquired sometime; the probability of seeing one pulse in unlimited time is 100 per cent. To run the most effective signal search, it is necessary to have the smallest possible value of T_H.

D. PERSISTENCE OF THE INDICATOR

Study of the graph, figure VI-3, shows that a search for pulse signals from scanning antennas may produce intervals up to and beyond 100 seconds, (1-2/3 minutes) between probable acquisitions for certain receiver and signal characteristics. For this reason an indicating device capable of storing information for a long time is required. In addition, a means for correlating the information on successive scans is required. To ensure complete acquisition, a printer-recorder is desirable.

Only for cases where there is positive knowledge that no scanning radars will be involved can there be justification for omitting the image storing type of indicator.

The optimum recorder type would be one which presents time along the length of a film or tape, frequency across the width of the tape and amplitude by intensity of the marks. A "falling raster" scan on a memory tube provides this type of presentation for shorter term storage.

E. RECEIVER DEAD-TIME

To maintain high acquisition probability, the sweep waveform of the receiver should provide as much active receiving time per sweep as possible. The maximum time would probably be provided by a continuously active receiver that used a triangular waveform with equal rise and fall time. When the receiver preselector and/or local oscillator contains electromagnetic tuning hysteresis effects in the local oscillator or preselector device may cause frequency shift between the upward scan and the downward scan. A linear sawtooth with rapid flyback is recommended because of the ease with which it can be produced, and because effects of hysteresis in magnetically tuned (YIG) filters and incremental inductance tuned circuits would be minimized. The usual need to supply a negative coercive force to erase residual magnetism will require a more or less trapezoidal waveform and a longer dwell between scans, thereby increasing receiver dead-time.

PROBABILITY OF INTERCEPT MATHEMATICAL ANALYSIS

A. UNIT COINCIDENCE PROBABILITY EQUATIONS

Equations for Probability of Intercept of a Pulsed Signal from a Scanning Antenna in a Single Random Sweep of a Panoramic Receiver with Sawtooth Sweep Pattern. The following equations for Unit Coincidence Probability, (UCP) were obtained from bibliography reference 1, Appendix VI-II, the terms being changed to agree with those used in the remainder of this report.

$$t_r < (T_s - t_s)$$

$$p^{(1)} = \frac{t_{ta}}{T_{ta}} \cdot \frac{t_{rs} + t_s}{T_s} \tag{VI-I-1}$$

$$= \frac{\alpha}{\beta} \cdot \frac{t_{rs} + t_s}{T_s}$$

where

$$(T_s - t_s) < t_{rs} < (T_s + t_s) \text{ or } [t_{rs} \approx T_s] \tag{VI-I-2}$$

$$p^{(1)} = \frac{t_{ta} + t_s}{T_{ta}}$$

where

$$(T_s + t_s) < t_{rs} < [(T_{ta} - t_{ta}) + (2 T_s - 2 t_s)] \tag{VI-I-3}$$

$$p^{(1)} = \frac{t_{rs} + t_{ta} - 2 T_s + 2 t_s}{T_{ta}}$$

where

$$[(T_{ta} - t_{ta}) + 2 (T_s - t_s)] < t_{rs} \tag{VI-I-4}$$

$$p^{(1)} = 1$$

In the equation above,

$$t_{rs} = k \frac{T_s}{B_r} (b_r + b_s) \tag{VI-I-5}$$

The first three regions above are further broken down and presented graphically in figure VI-I-1.

The following derivations result in general equations, which, when the limiting conditions are imposed, result in the above equations.
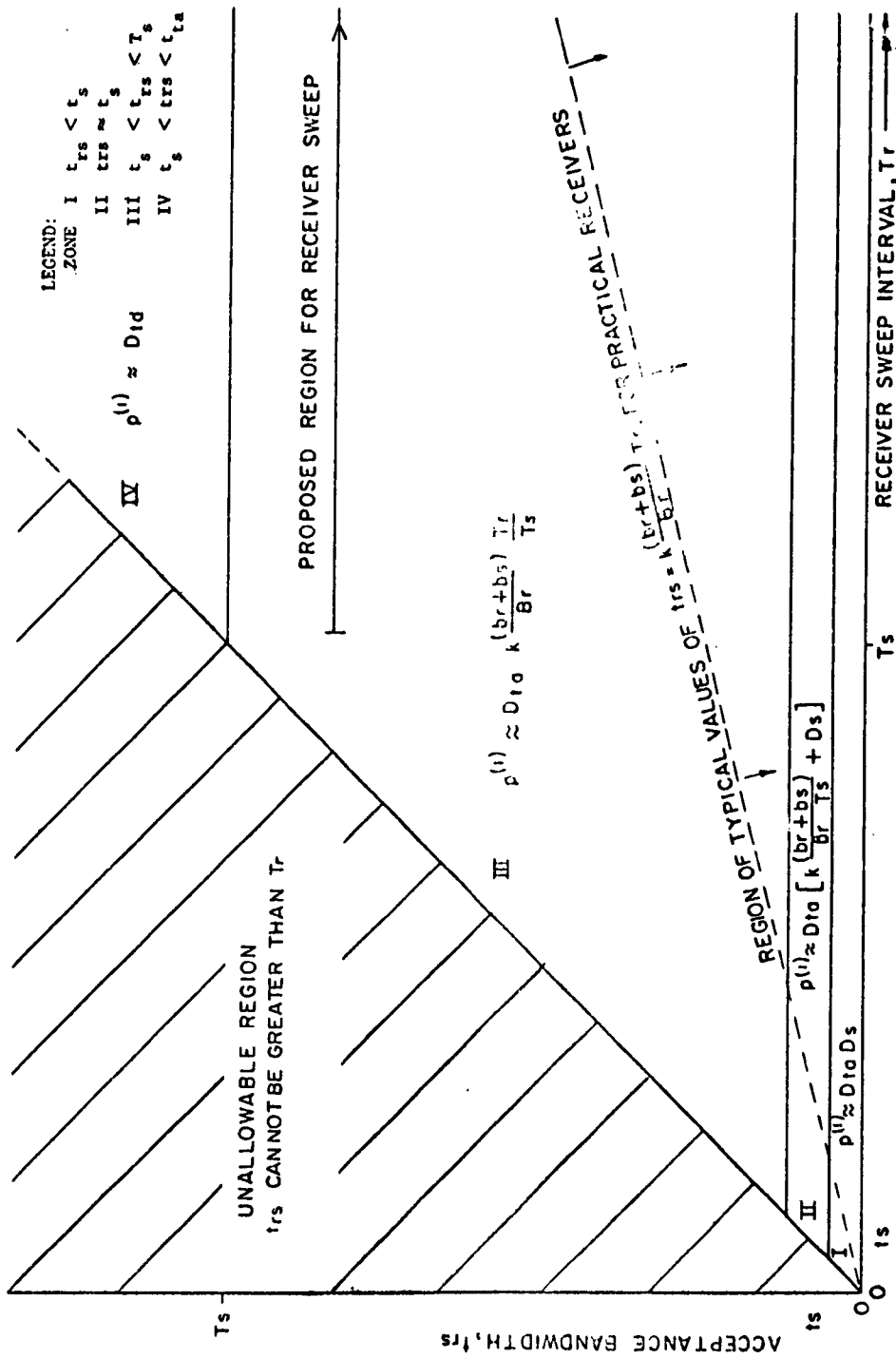
Figure VI-I-1. Regions for Unit Coincidence Probability vs. Receiver and Signal Parameters

B. DERIVATION OF THE PROBABILITY EQUATIONS

1. Probability of Intercept from a Fixed Radiator

To find the probability of acquiring a signal on one sweep, we must determine the ratio of successful pulse start time to the total time in the sweep.

Refer to figure VI-I-2(a) which shows a typical pulsed signal and a typical receiver sweep. From this drawing, we may determine the probability of seeing a pulse from a signal generator connected directly to the receiver.

Figure VI-I-2(b) shows the location of signal pulses at the two time limits which will be assumed to produce some acquisition by the receiver by infinitesimal overlap of frequency/time.

The total pulse-starting time of pulses which will be considered successfully acquired is

$$\Sigma t = t_s + t_r + t_1 \qquad\qquad\qquad (VI-I-6)$$

by solution of similar triangles (refer to figure VI-I-2(a)) it can be shown that

$$t_r = kT_r \; \frac{b_r}{B_r} \qquad\qquad\qquad (VI-I-7)$$

$$t_1 = kT_r \; \frac{b_s}{B_r} \qquad\qquad\qquad (VI-I-8)$$

then $\qquad \Sigma t = t_s + k \; \dfrac{T_r}{B_r} \; (b_r + b_s) \qquad\qquad (VI-I-9)$

let us define the total time within which any part of the signal spectrum may coincide with the receiver acceptance band as t_{rs}:

$$t_{rs} = t_r + t_1$$

substituting from equations (VI-I-7) and(VI-I-8)

$$t_{rs} = k \; \frac{T_r}{B_r} \; (b_r + b_s) \qquad\qquad\qquad (VI-I-9a)$$

and if one pulse occurs per sweep period, the probability of acquiring a signal on a single random sweep (P_1) is: equal to the ratio of possible successful pulse starting times per sweep to the total time per sweep

a.
RELATIONSHIP OF RECEIVER
& SIGNAL PARAMETERS

b.
LIMITS OF ASSUMED COINCIDENCE

Figure VI-I-2. Typical Pulsed Signal on Panoramic Receiver.

$$P_1 = \frac{\Sigma_t}{T_r} = \frac{t_s + k \frac{T_r}{B_r} (b_r + b_s)}{T_r} \qquad \text{(VI-I-10)}$$

$$= K \frac{(b_r + b_s)}{B_r} + \frac{t_s}{T_r}$$

Where $T_r \neq T_s$, there will be T_r/T_s pulses per sweep interval and

$$p^{(1)} = P_1 \frac{T_r}{T_s} = \left[K \frac{(b_r + b_s)}{B_r} + \frac{t_s}{T_r} \right] \frac{T_r}{t_s}$$

$$= K \frac{(b_r + b_s)}{B_r} \cdot \frac{T_r}{T_s} + D_s \qquad \text{(VI-I-11)}$$

where $D_s = t_s/T_s$

when $p^{(1)} = 1$, one pulse will be received on every sweep and the fastest sweep which will permit receipt of one pulse per sweep can be determined. Setting equation (VI-I-11) equal to 1, that sweep speed f_{rm} can be found.

$$p^{(1)} = K \frac{(b_r + b_s)}{B_r} \frac{T_r}{T_s} + D_s = 1$$

$$k \left(\frac{\frac{T_s (1-D_s)}{b_r + b_s}}{B_r} \right) = T_{rm}$$

$$f_{rm} = \frac{1}{T_{rm}} = \frac{K \left(\frac{b_r + b_s}{B_r} \right)}{(1 - D_s)} f_s \qquad \text{(VI-I-12)}$$

It may be noted that in the case of a CW signal, $D_s = 1$, the equation is indeterminate. For CW, an acquisition is obtained on every sweep regardless of sweep-rate.

For the usual radar signal,

$$D_s \ll 1, \text{ and } (1 - D_s) \approx 1$$

and

$$f_{rm} \approx K \ \frac{(b_r + b_s)}{B_r} \ f_s \qquad \text{(VI-I-13)}$$

2. <u>Probability of Intercept of Signals from a Rotating Antenna</u>

Where the signal eminates from a rotating antenna there will be T_{ta}/T_s pulses emitted in a full rotation of the antenna but only α/θ x T_{ta}/T_s pulses will be emitted during the time of each antenna rotation that the antenna beams a signal toward the receiver reducing the vaule of $p^{(1)}$ proportionally. Unit probability for signals from a scanned antenna becomes, from equation (VI-I-11)

$$P_a^{(1)} = \frac{\alpha}{\theta} \ p^{(1)} = \frac{\alpha}{\theta} \left[K \ \frac{(b_r + b_s)}{B_r} \ \frac{T_r}{T_s} + D_s \right] \qquad \text{(VI-I-14)}$$

In the case of radiations from a scanned antenna it is almost always certain that at least one pulse will be transmitted toward the receiver in the time the beam covers the receiver on each scan. We might also make the assumption that many radar sets will emit two or three pulses in the time the beam of its antenna covers the receiver antenna.

We are therefore ensured of one or more pulse interceptions on each sweep of the receiver, if

$$t_{rs} = T_{ta} - t_{ta} \qquad \text{(VI-I-15)}$$

In this case the signal bandwidth becomes an insignificant part of the problem, and from equation (VI-I-9a):

$$t_{rs} \approx K \ \frac{b_r}{B_r} \ T_r \qquad \text{(VI-I-16)}$$

It can also be shown that

$$t_a = \frac{\alpha}{\theta} \ T_{ta} \qquad \text{(VI-I-17)}$$

The maximum receiver sweep frequency for acquiring one pulse from a scanning antenna is found by substitution in equation (VI-I-15) setting

$$p^{(1)} = 1:$$

VI-I-6

$$k \, \frac{b_r}{B_r} \, T_r = T_{ta} \, (1 - \frac{\alpha}{\theta})$$

for a typical radar

$$\alpha/\theta \le 1/100 \text{ and } (1 - \alpha/\theta) \approx 1$$

$$\therefore T_r \approx T_{ta} // k \, b_r/B_r \qquad\qquad T_r = T_{ta}/k \, \frac{b_r}{B_r}$$

$$f_{rm} = \frac{1}{T_r} \approx k \, \frac{b_r}{B_r} \, f_{ta} \qquad\qquad \text{(VI-I-18)}$$

for typical antennas, this results in sweep periods (and, consequently, minimum acquisition intervals) in the order of 10-to-100 minutes or more.

APPENDIX VI-II

THE ACQUISITION INTERVAL

A. DISCUSSION

The important aspect of the investigation is the determination of the acquisition interval, which sets the time a panoramic receiver must look to be sure of having acquired every continuously radiated pulse signal in the environment being investigated. The determination also gives an idea of the length of time an intermittent signal must persist to be sure of being acquired.

The exact determination of acquisition interval by mathematical approach is extremely difficult and apparently has been carried out only for specific conditions applying to individual programs. The following charts and equations will show some easily determined intervals for special cases and will attempt to demonstrate the trend for other cases.

1. Continuous signals, CW(AΦ), FM, AM, will be received on every sweep of the receiver making $T_H = T_r$.

2. Pulses where there is a synchronous relationship between the pulse-repetition frequency of the signal, $1/T_s$, and the sweep frequency of the receiver, $1/T_r$, fall into two classes.

a. Where $T_s = n T_r$, n being any integer ≥ 1, a pulse, once having been acquired on a given sweep, will be acquired on every subsequent sweep. Again, $T_H = T_r$. If the exact relation $T_s = nT_r$ is maintained, the signal will never be acquired; $T_H = \infty$

b. Where $nT_s = T_r$, m being any integer > 1, a pulse acquired on one sweep will be acquired again every m sweeps. As before, if the pulse is missed, and the exact relation $mT_s = T_r$ holds, the pulse will never be acquired. Again, $T_H = \infty$

c. To avoid the chance of missing signals due to synchronism, the sweep start can be jittered, in which case the acquisition interval, T_H, becomes a function of the jittering signal.

3. When the dwell time t_{rs}, of the receiver is greater than the pulse interval of the signal, one or more pulses will be acquired on each sweep of the receiver. For the purpose of this program, a single acquisition will be considered, since the operator has noted the presence of only one signal even though more than one pulse of that signal occurred. The acquisition interval, T_H is again equal to T_r.

For this condition, a maximum sweep frequency has been determined in Appendix VI-I.

Equation (VI-I-13), applies to the case of a signal radiated from a stationary antenna

$$f_{rm} \approx k \ \frac{b_r + b_s}{B_r} \ f_s$$

and equation (VI-I-18), applies to the case of a signal radiated from a rotating antenna having a narrow beam

$$f_{rma} \approx k \frac{b_r}{B_r} f_{ta}$$

4. For other signals, where $T_s \neq nT_s \neq T_r$; and

$$T_r < \frac{B_r}{k(b_r + b_s)} \quad T_s \quad \text{or} \quad T_r < \frac{B_r}{K b_r} \quad T_{ta} \quad \text{(whichever applies)}$$

writing a general equation for Acquisition Interval becomes extremely difficult.

a. One fact can be stated with certainty. Since the receiver tunes each signal frequency only once per sweep, the acquisition interval can never be shorter than the period of a receiver scan, T_r:

$$T_H \overset{<}{=} T_r \tag{VI-II-1}$$

The graph, Figure VI-II-1, represents the minimum possible acquisition interval for known values of sweep rate, $f_r = 1/T_r$

b. Examination shows that for the cases where the signal pulse is related to the receiver sweep in a rational manner, $nT_r = mT_s$, or $T_r = m/n \, T_s$, where m and n are both integers, then signal pulses (of infinitesimal length) will be received once each n sweeps, and the acquisition interval, $T_H \overset{<}{=} n \, T_r$. e.g. $T_r = m \, T_s$

    1 $T_r = 2 \, T_s$ an acquisition occurs on each sweep

    2 $T_r = 1 \, T_s$, or $T_s = T_r/2$ an acquisition occurs on each second sweep

    3 $T_r = T_s$, or $T_s = T_r/3$ an acquisition occurs on each third sweep

Or, take the case where 463 $T_r = 230T_s$

which makes $2T_r > T_s$. However, the hits occur, not once every second sweep as it would when $2T_r = T_s$, but rather on every 230th sweep. This same pattern holds throughout; changing the ratio T_r/T_s by an infinitesimal amount may make a large difference in acquisition interval.

To demonstrate the inability to develop a simple general equation, we may plot three points on a curve, e.g.

    2 $T_r = T_s$ acquisition interval   2 T_r

    4 $T_r = T_s$ acquisition interval   4 T_r

    6 $T_r = T_s$ acquisition interval   6 T_r

If we now add points between, for example:
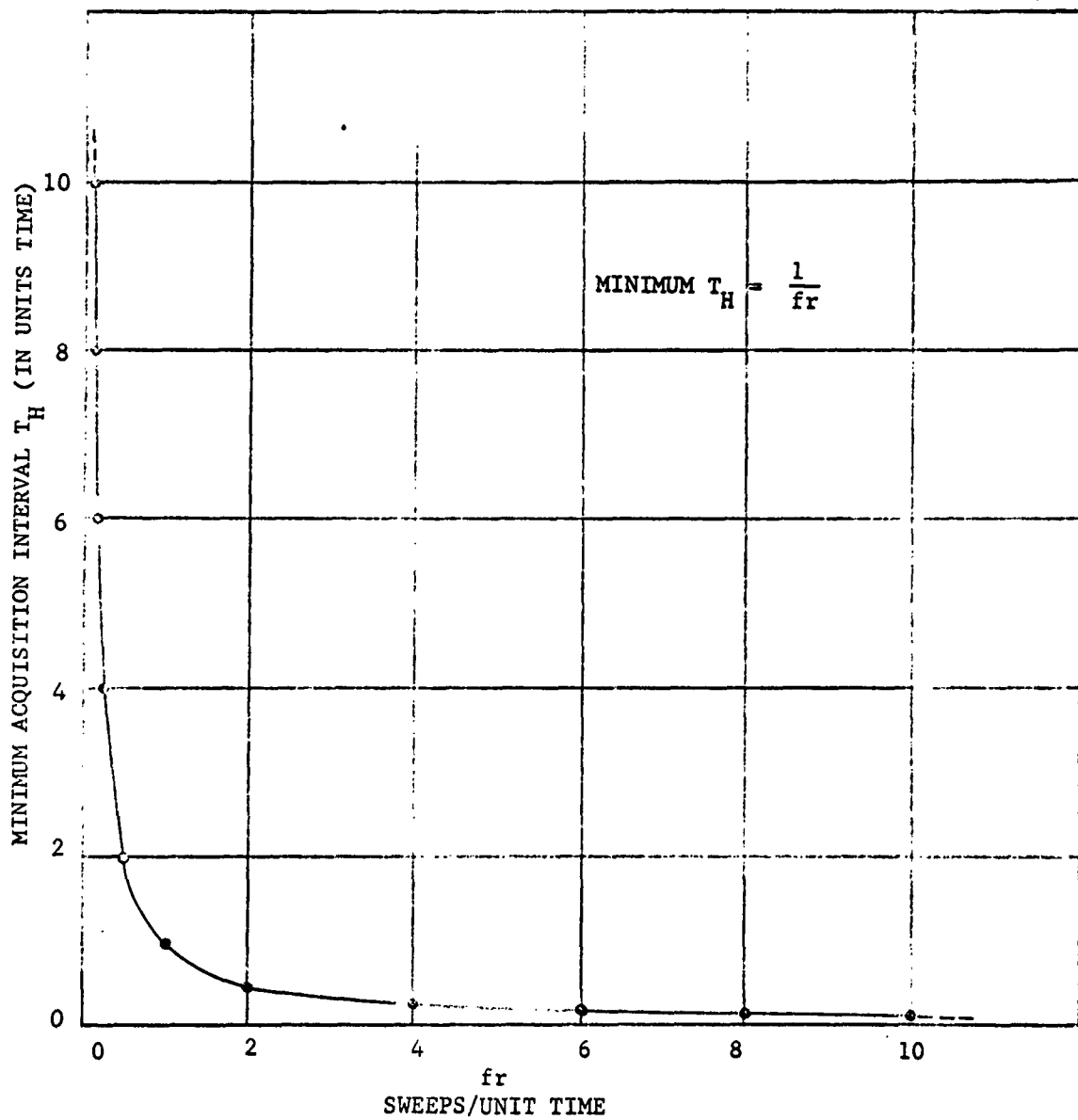
Figure VI-II-1. Minimum Possible Acquisition Interval

$$7 \, T_r = 3 \, T_s \qquad T_r = 3/7 \, T_s \qquad \text{acquisition interval} = 7 \, T_r$$

$$9 \, T_r = 4 \, T_s \qquad T_r = 4/9 \qquad \text{acquisition interval} = 9 \, T_r$$

$$2 \, T_r = 5 \, T_s \qquad T_r = 5/2 \, T_s \qquad \text{acquisition interval} = 2 \, T_r$$

these new points, though between the points previously plotted, do not fall on the old curve.
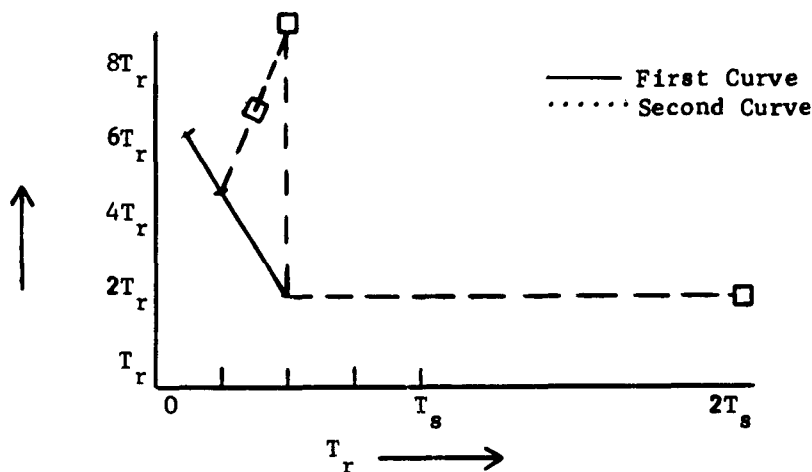
Figure VI-II-2.

New points between these points may fall above or below the new dotted line. As finer division takes place, the dispersion will increase. It is seen that no curve for acquisition interval, based on the equation $nT_r = T_s$, can be plotted.

The information developed up to this point has enabled us to find the minimum possible acquisition interval ($T_H = T_r$), and some intervals for special cases, e.g., $T_H = nT_r$ when $nT_r = mT_s$, n and m being positive integers.

The bandwidth characteristic of the receiver sets another limit.

In Appendix VI-I, equation (VI-I-13) was

$$f_{rm} = k \, \frac{b_r + b_s}{B_r} \, f_s$$

which can be rewritten

$$T_{r_{min}} = \frac{B_r + T_s}{k(b_r + b_s)} \qquad\qquad (VI\text{-}II\text{-}2)$$

This investigation has established certain limitations that are valid for relating the pulse rate, the sweep rate and the acquisition interval. These limits are plotted for a general case in Figure VI-II-3 where an assymptote for minimum acquisition interval of nonsynchronous pulses is less than infinity.

As stated above,

$$T_H = n\, T_r \text{ when } T_r = \frac{m}{n}\, T_s \qquad\qquad (VI\text{-}II\text{-}3)$$

From equation (VI-I-11), the unit coincidence probability for a signal with infinitessimal pulses is

$$P_i^{(1)} = k\, \frac{(b_r + b_s)}{B_r}\, \frac{T_r}{T_s}$$

for signal with finite pulses the probability improves. In the equation

$$P_f^{(1)} = k\, \frac{(b_r + b_s)}{B_r}\, \frac{T_r}{T_s} \text{ let } D_r = k\, \frac{b_r + b_s}{B_r}$$

then

$$\frac{P_{-f}^{(1)}}{P_i^{(1)}} = \frac{\dfrac{D_r T_r}{T_s} + D_s}{D_r\, \dfrac{T_r}{T_s}} = 1 + \frac{D_s T_s}{D_r T_r} \qquad\qquad (VI\text{-}II\text{-}4)$$

but $D_s = \dfrac{t_s}{T_s}$ and it can be shown that $D_r = \dfrac{t_{rs}}{T_r}$

enabling equation (VI-II-4) to be rewritten

$$\frac{P_f^{(1)}}{P_i^{(1)}} = 1 + \frac{D_s\, T_s}{t_{rs}} = 1 + \frac{t_s}{t_{rs}} \qquad\qquad (VI\text{-}II\text{-}5)$$

with this improved probability it may be assumed that the probability will be reduced by the reciprocal ratio in which case

$$T_{Hf} = n\, T_r \times \frac{1}{1 + \dfrac{t_s}{t_{rs}}}$$
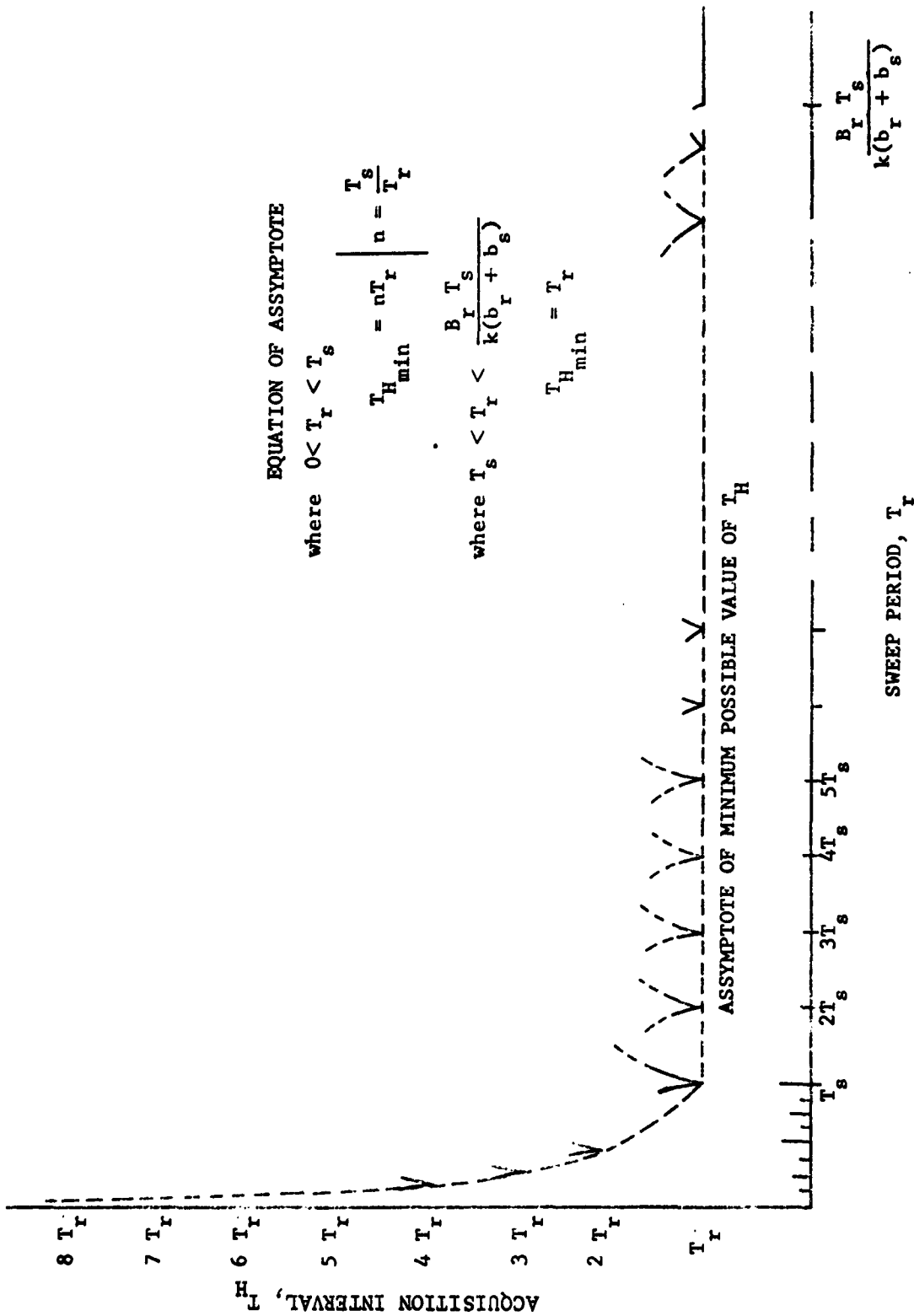
Figure VI-II-3. Acquisition Intervals

EQUATION OF ASSYMPTOTE

where $0 < T_r < T_s$

$$T_{H_{min}} = nT_r \;\Big|\; n = \frac{T_s}{T_r}$$

where $T_s < T_r < \dfrac{B_r T_s}{k(b_r + b_s)}$

$$T_{H_{min}} = T_r$$

ASSYMPTOTE OF MINIMUM POSSIBLE VALUE OF T_H

SWEEP PERIOD, T_r

ACQUISITION INTERVAL, T_H

$$\frac{B_r T_s}{k(b_r + b_s)}$$

In the case of the usual radar pulse the receiver will usually be operating in a manner that results in ratios $\dfrac{t_s}{t_{rs}}$ in the order of 1/2 or less and the Acquisition Interval might be expected to be about 1/3 less than that determined from the assumption that $T_H = nT_r$ when $mT_s = nT_r$.

B. CONCLUSIONS

This analysis has shown that certain limiting conditions to guide the receiver designer may be deduced from an analysis of Acquisition Interval and Unit Coincidence Probability, but no accurate simple general equation is possible. Figure VI-II-3 may aid in visualizing the limits determined from this analysis.

C. REFERENCES

Information for Appendix VI was taken from each of the below-named soruces. The two sources in turn list some 27 other references, many of which treat special aspects of the total problem.

1. P.H. Endlow, Jr., "Some Techniques for the Analysis of Intercept Probability in Reconnaissance Receivers", Technical Report #516-1, Systems Techniques Laboratory, Stanford Electronic Laboratories, Stanford U., Stanford, Calif., June 1959. (Unclassified) (ASTIA Code No. AD 220387).

2. H. W. Batten, R. A. Jorgensen, A. B. Mac Nee, and W. W. Peterson, "The Response of a Panoramic Receiver to CW and Pulse Signals," Proc. of IRE, vol. 42, pp. 948-956, June 1954.

REFERENCES

Information for Appendix VI was taken from each of the below-named sources. The two sources in turn list some 27 other references, many of which treat special aspects of the total problem.

1. P.H. Endlow, Jr., "Some Techniques for the Analysis of Intercept Probability in Reconnaissance Receivers, "Technical Report #516-1, Systems Techniques Laboratory, Stanford Electronic Laboratories, Stanford U., Stanford, Calif., June 1959. (Unclassified) (ASTIA Code No. AD 220387).

2. H. W. Batten, R. A. Jorgensen, A. B. MacNee, and W. W. Peterson, "The Response of a Panoramic Receiver to CW and Pulse Signals", Proc. of IRE, vol. 42, pp. 948-956, June 1954.

A. GENERAL

Since the receiver front-end approach has been chosen (Appendix V) and the detection probability calculated (Appendix VI), unit specifications can now be written for the preselectors, the local oscillators and the IF amplifiers.

The results of the following discussions are tabulated in Table V, and in Figure 2 of the text.

B. PRESELECTOR SPECIFICATIONS

Before preselector specifications can be chosen, it is necessary to know the IF bandwidth of the receiver as well as the IF center frequency. The IF bandwidth will be equal to the minimum bandwidth acceptable from the preselector; the IF center frequency will determine the maximum preselector bandwidth for a given image rejection or local-oscillator radiation.

To allow for some local-oscillator instability, as well as local oscillator-preselector tracking error, the preselector bandwidth should be wider than the IF bandwidth by approximately 20 percent. Since most preselectors change bandwidth with center frequency, it is probably optimum to adjust the mean bandwidth to be 20 percent greater than the IF bandwidth.

The preselector bandwidth should not be more than 20 percent greater than the IF, if excessive demands are not to be brought on the filtering characteristics of the preselector. Aside from accepting only the signal frequency, the preselector must also; (1) isolate the local oscillator from the antenna so that radiation does not occur, and (2) prevent a response at the image frequency.

A typical preselector band pass is illustrated in Figure VII-1.

Local-oscillator rejection of -67 dbm is called out in RFI specifications. Up to approximately 500 Mc, RF stages will be utilized between preselector stages. The reverse isolation of these stages, approximately 30 db per stage, is more than adequate to reduce the radiated local-oscillator power to -67 dbm.

Above 500 Mc, however, no RF preamplification may be required. The preselector must therefore attenuate the local-oscillator signal. As shown in Appendix V, four-section preselectors are required to provide adequate image rejection. Again, from

$$\beta = \frac{2\Delta f}{4\sqrt{\sigma}} \; , \qquad\qquad\qquad \text{(VII-1)}$$

$$30 = \frac{190}{4\sqrt{\sigma}} \qquad\qquad\qquad \text{(VII-2)}$$
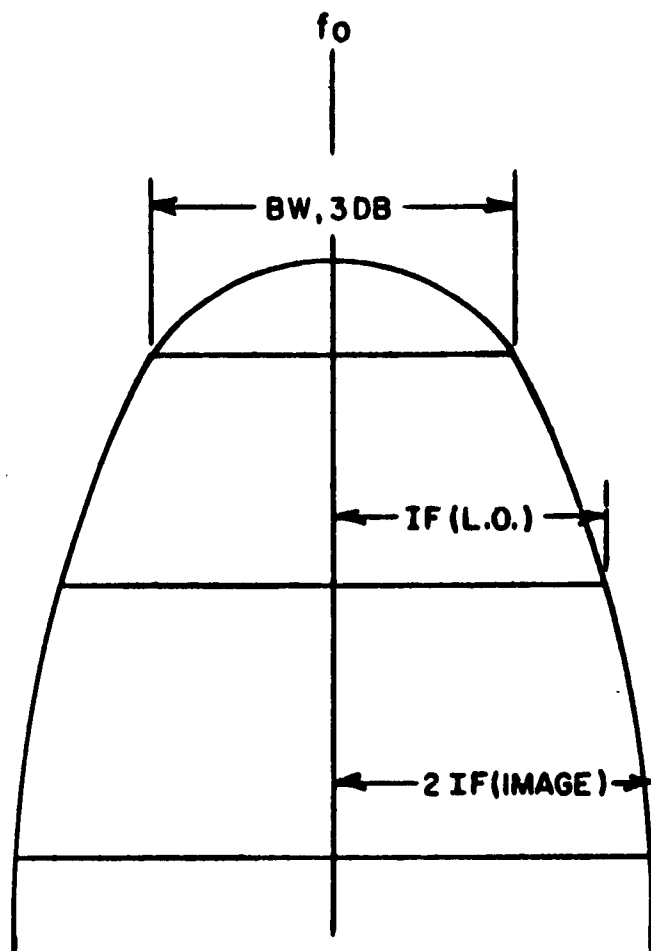
Figure VII-1.Typical Preselector Bandpass

$$\sigma = 6.34 \qquad\qquad\qquad \text{(VII-3)}$$

$$\sigma \approx 1600 \cong -64 \text{ dbm.} \qquad\qquad \text{(VII-4)}$$

Thus, 64 db of local-oscillator rejection is provided by the preselector. However, an additional 20 db of local-oscillator isolation is achieved with the use of a balanced mixer. Therefore, if we assume that there is a local-oscillator level of 1 mw, local-oscillator radiated power would be -84 dbm.

C. LOCAL-OSCILLATOR SPECIFICATIONS

It is the intent of the present program to investigate and propose a receiver comprised entirely of solid-state components. The local oscillator must therefore be solid state and must provide the frequency coverage specified in Figure 2. Actually, the frequency range should be slightly greater than the frequency ranges specified, since signal-band overlap is desirable.

The power output of the local oscillator must be sufficient to drive the mixer; for example, a crystal mixer requires approximately 1 mw and a tunnel diode mixer requires on the order of -20 dbm.

Local-oscillator stability should be on the order of 1 part in 10^4. When the local oscillator is sweeping, the ends of the swept band should not change more than 1 part in 10^4, and when it is not sweeping, the short time stability should be 1 part in 10^4. When AFC is utilized, stabilities of 1 part in 10^6 or better should be produced.

D. IF AMPLIFIER SPECIFICATIONS

Local oscillator and IF amplifier center frequencies were selected previously, as indicated in Figure 2. IF amplifier bandwidth has yet to be determined.

Detection probability is seen to increase with IF bandwidth in a swept receiver, i.e. $P_D \sim b/B$.

where: P_D = detection probability

  b = IF bandwidth

  B = IF center frequency

Therefore, b should be as large as possible while still providing some coarse resolution and good sensitivity. Approximately -90-dbm sensitivity can be achieved with a b/B ratio of 1/50.

However, since a sector scan of as little as 5 percent will be utilized, a narrow bandwidth is required to achieve high resolution. A b/B ratio of approximately 1/4000 is therefore required. Other bandwidths should be available between these limits to provide either continuous bandwidth variation or stepped variations at b/B ratios

of 1/50, 1/150, 1/450, 1/1350, and 1/4000.

The IF amplifier characteristics will be lin-log. The logarithmic characteristics will enable strong and weak signals to be seen without a change of the manual gain setting. The IF characteristic will be linear for the first 10 or 20 db of gain in order to get the signal out of the noise rapidly.

The IF amplifier gain shall be sufficient in each case to amplify the noise to a detectable level.

APPENDIX VIII

R-F FRONT END: 100 KC - 250 MC

A. GENERAL

Since $f = 1/2\pi\sqrt{LC}$, where f = center frequency, L = inductance and C = capacitance, there are only two variables which can be varied in order to tune preselector and oscillator circuits over an octave frequency range. An investigation has been made into the availability of electrically tuned capacitors and inductors in the range of 100 kc to 250 Mc, and the results of this investigation are presented in the paragraphs which follow.

B. ELECTRICALLY TUNED CAPACITORS

1. Theory

Of the number of these components that are on the market, those produced by P.S.I. exhibit optimum characteristics for incorporation in the receiver under consideration. This is because P.S.I. has a line of varicaps which have a higher Q than is generally found in this type of device. A circuit Q of 50 has been proposed for the preselector tuned circuits, since this is believed to give a bandwidth which will be wide enough for the acquisition of all signals in the frequency range. In the standard fixed or manually tuned circuit the capacitor usually has such a high Q (on the order of a few thousand) that it has very little effect on the Q of the circuit. The inductor with its inherent wire resistance, and the accompanying circuitry, are the determining factors of the circuit Q. It is difficult to achieve a high Q in circuits which use an electrically tuned capacitor because the capacitor lowers the overall circuit Q. Typical Figure of Merit (Q) versus bias-voltage characteristics at 25°C for the high-Q varicap are given in Figure VIII-1.

By picking the high-Q type of voltage-controlled capacitor, the range of capacitance change has been limited. With the lower-Q type of voltage-controlled capacitors, ranges of approximately 10:1 can be received; and with the high-Q variety, a maximum range of capacitance change (C_{2v}/C_{100v}) is 5.2:1. This range should be enough to achieve the desired octave tuning range. Typical capacitance versus bias-voltage characteristics at 25°C are given in Figure VIII-2. As has already been mentioned:

$$f = \frac{1}{2\pi\sqrt{LC}}$$

For an octave range, the sweep will be 2f,
or

$$2f = \frac{1}{2\pi\sqrt{LC}} \tag{VIII-1}$$

$$f = \frac{1}{4\pi\sqrt{LC}} \tag{VIII-2}$$

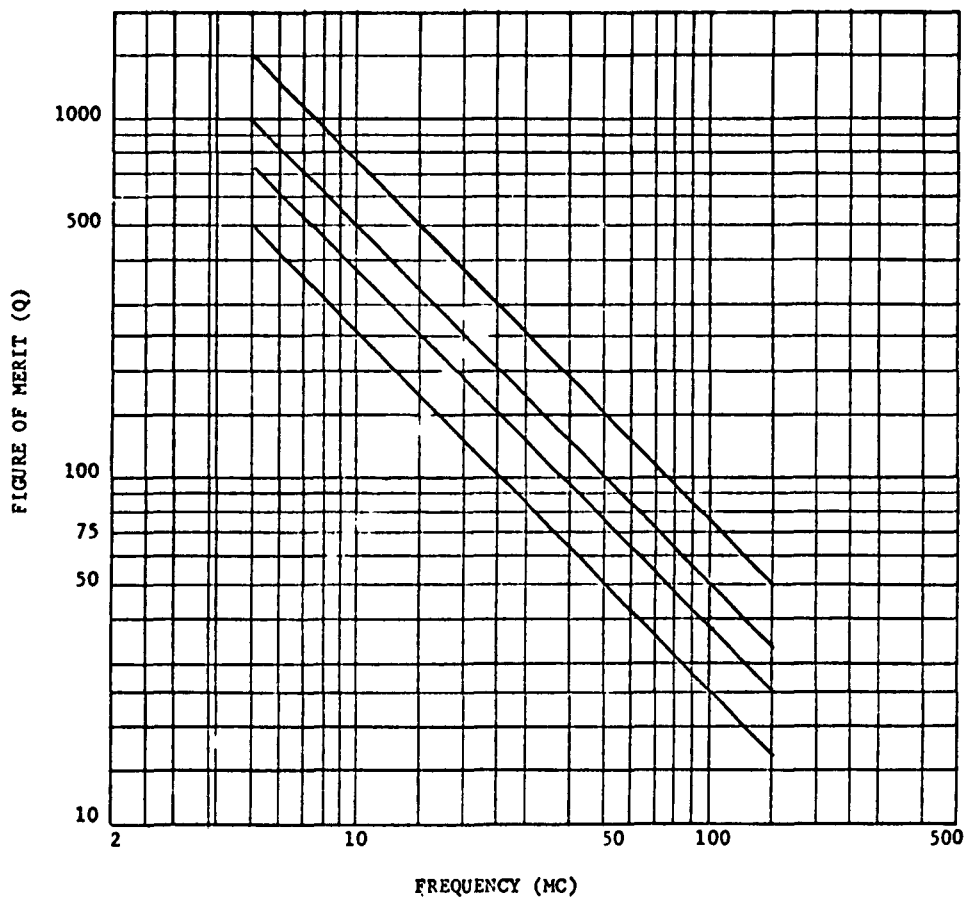| SILICON HIGH Q VARICAP	Typical Figure of Merit (Q) Versus Bias Voltage Characteristics at 25°C	Technical Data 10100-1 Rev. 2
	Applies to: PC-107; PC-112: PC-113; PC-114; PC-115: PC-116; PC-117; PC-122; PC-132; PC-133; PC-134: PC-135: PC-136; and PC-137.	Dwn by: _____ App by: _____ Rel by: _____ Date: 9-13-6C
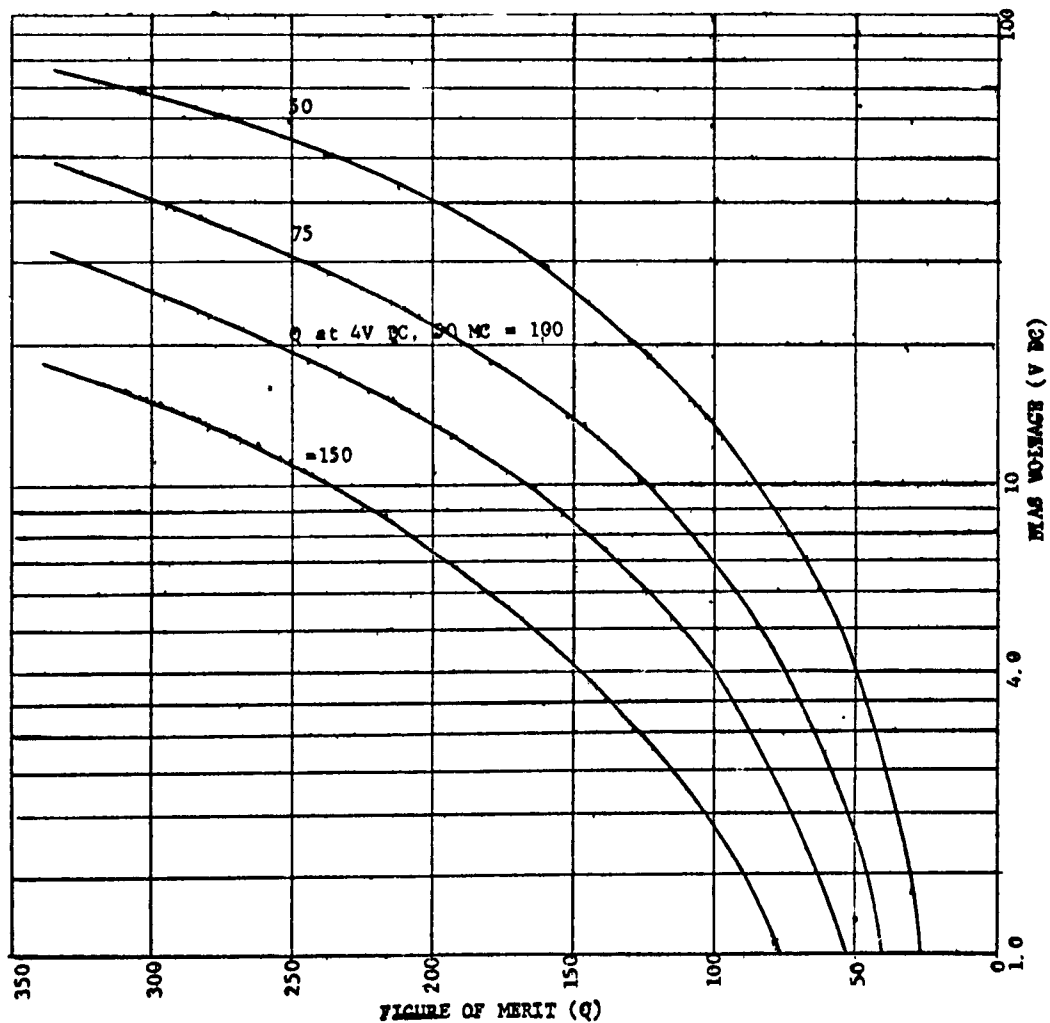
Figure VIII-1. Typical Figure of Merit (C) Versus Bias Voltage Characteristics a. 2°C

VIII-2

SILICON HIGH Q VARICAP	Typical Capacitance Versus Bias Voltage Characteristics at 25°C	Technical Data 10100 - 5 Rev 1
	Applies to: PC-107; PC-112; PC-113;PC-114; PC-115; PC-116; PC-117;PC-122; PC-132; PC-133; PC-134;PC-135; PC-136 and PC-137	Dwn by: ——— App by: ——— Rel by: ——— Date: 9-19-60
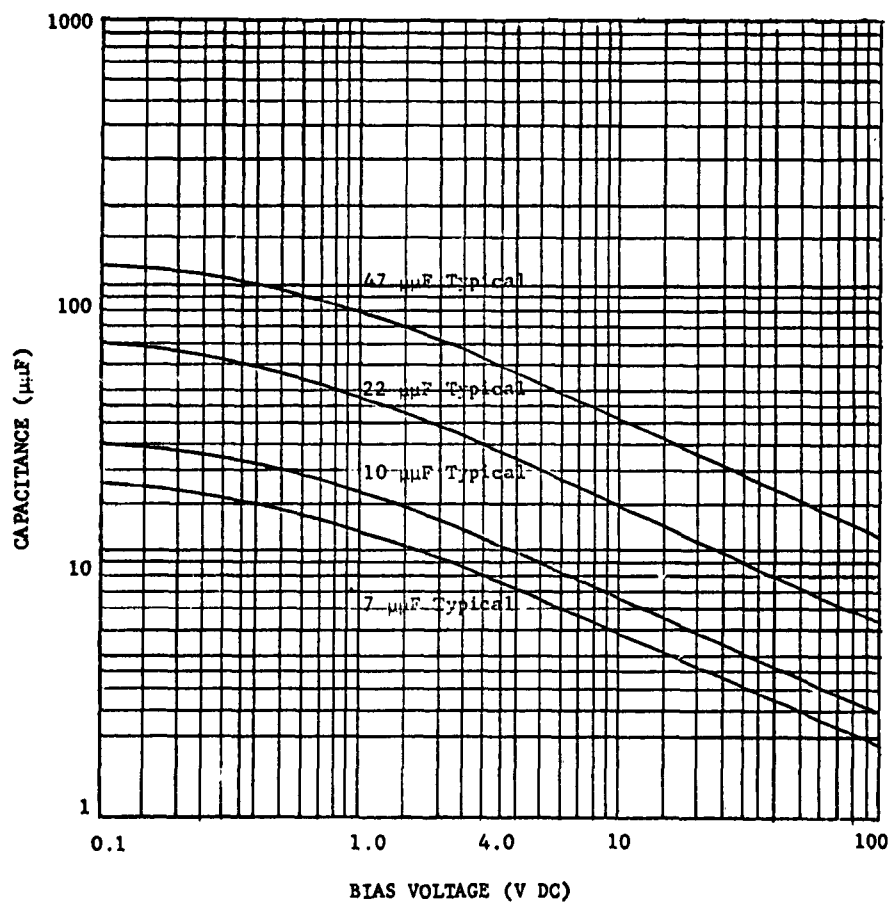
Figure VIII-2. Typical Capacitance Versus Bias Voltage Characteristics at 25°C

therefore

$$f = \frac{1}{2\pi\sqrt{L(4C)}}$$
(VIII-3)

which states that a capacitance change of 4:1 will be enough to tune the frequency over an octave range. However, since the "C" of the equation is the total circuit capacitance, including stray capacitance, output and input transistor capacitance, and the varicap capacitance, the capacitance change of the varicap must be greater than 4:1 to make certain that the total circuit capacitance change is 4:1.

To sweep the entire octave range, a 100V saw-tooth voltage is required for the varicap. The capacitance variation with bias voltage is not a linear one, nor would the frequency variation vs. bias be linear, as is seen in Figure VIII-3. The drive will therefore have to be compensated so that a linear frequency vs. bias characteristic can be obtained.

Along with the problem of a linear frequency vs. bias characteristic, there is the additional requirement that the tuned circuits used in the preselector will have to track the local oscillator. Since the bandwidth requirement is 1/50th of the center frequency, the preselector and the local oscillator will have to track within a ± 1/250 of the center frequency. Due to the fact that the entire range of capacitance is being used to sweep the desired octave range, trimmer capacitors can not be used to compensate for tracking errors. The required compensation must therefore be handled by means of the sweep circuit that will be used. All the varicaps will have the same basic bias, but each will have compensation to adjust for tracking.

The maximum frequency at which these high-Q varicaps can operate, with a minimum Q of 60, is approximately 100 Mc, as is seen in Figure VIII-4.

2. Feasibility Breadboard

A circuit was developed to receive signals in the 8- to 16-Mc range (Band 7 in Figure 2). Four stages of single-tuned preselection are used along with a Hartley Oscillator. The 9.5 to 17.5 range for the oscillator was achieved and a compensation network necessary to make the four RF stages track the local oscillator while separated by the IF (1.5 Mc) was built. A circuit diagram is shown in Figure VIII-5. A photograph of the breadboard circuit is shown in Figure 4 of the text.

Upon considering future temperature requirements of the circuit, silicon transistors were chosen. A transistor was required with low input and especially low output capacitance because of the need to keep all capacitance other than that of the "varicap" to a minimum in order that an octave range might be achieved. The reason that low transistor output capacitance is especially important, is that this total capacitance is across the tank circuit, while the input capacitance is reduced greatly due to being tapped down on the transformer used in
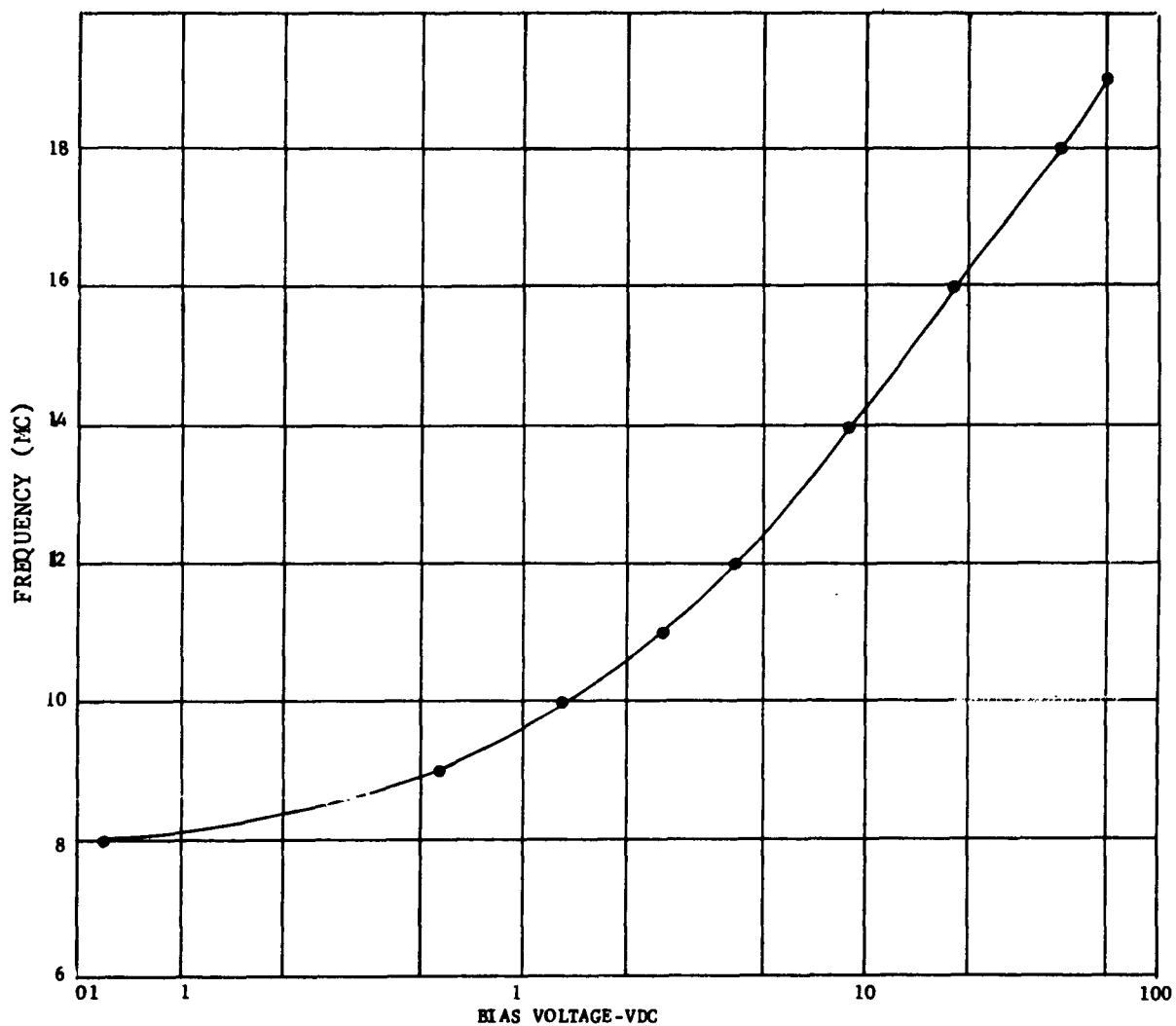
Figure VIII-3. Measured Frequency Variation Versus
Bias Voltage for a Two Stage Amplifier

SILICON HIGH Q VARICAP	Typical Figure of Merit (Q) Versus Frequency Characteristics at 25°C	Technical Data 10100-3 Rev. 2
	Applies to: PC-107; PC-112; PC-113; PC-114; PC-115; PC-116; PC-117; PC-122; PC-132; PC-133; PC-134; PC-135; PC-136 and PC-137	Dwn by: ____ App by: ____ Rel by: ____ Date: 9-13-60

Figure VIII-4.   Typical Figure of Merit (Q) Versus
Frequency Characteristics at 25°C

VIII-6
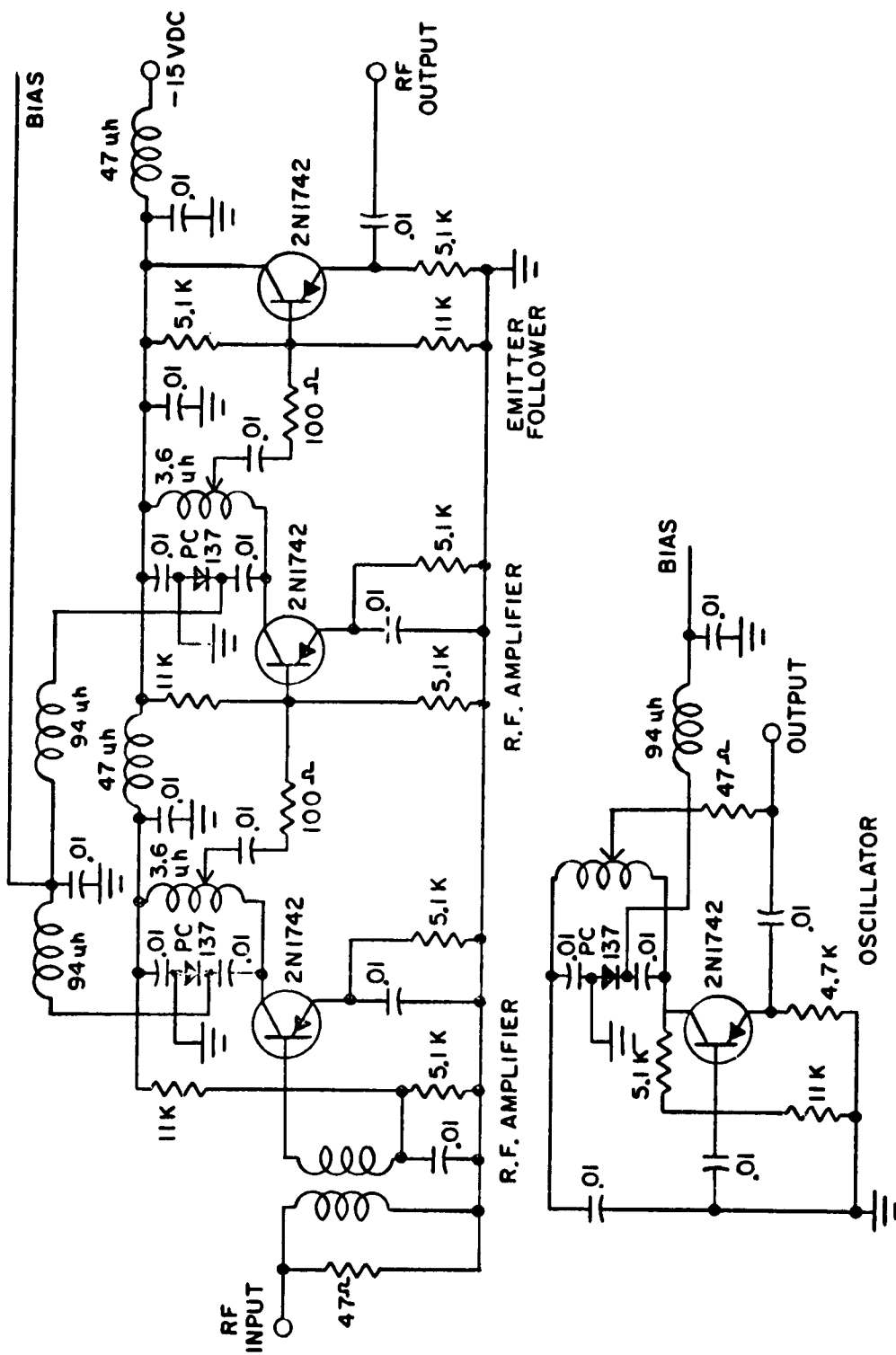
Figure VIII-5. Varicap-Tuned Preselector and Oscillator

1101466

the tank circuit. A 2N918 was chosen because of its low capacitance and low noise figure.

   a. Tracking

      (1) Two Point Tracking

When the bias voltage vs. capacitance curves (figures VIII-6 to VIII-11) were obtained for the "varicaps" to be utilized, it appeared as if a dc offset on the bias at the high capacitance end would be sufficient to make them have a common tracking point at 8 Mc. This dc offset has very little effect on the capacitance at the low capacitance end since the slope of capacitance vs. bias approaches 0 in this region. In order to be able to have a common point at this low capacitance end, trimmer capacitors were needed. As was stated in the first quarterly report, the use of trimmer capacitors with the "varicaps" having a capacitance ratio of C_{1v} to $C_{50v} \geq 4$ would make it difficult to achieve an octave range. A "varicap" with a capacitance ratio of C_{2v} to $C_{10v} \geq 5.2$ was installed in each circuit to make it possible to use a trimmer capacitor.

The problem of this means of tracking soon became evident. Due to a difference in the rate of change of capacitance versus bias for each "varicap", the circuit failed to track as well as was desired in the mid-frequency range.

      (2) Three-point Tracking

From the results received using two point tracking, another point was needed in the mid-frequency area to make the circuit track in this region. This was achieved by the use of a gain control for each varicap in order to shift the capacitance versus bias curve of each varicap to a common point in the mid-capacitance region. By adjusting the gain control for the mid-frequency range, the dc offset for the low frequency range and the trimmer capacitors for the high frequency range, tracking was achieved over the octave range.

B. ELECTRICALLY TUNED INDUCTORS

   1. Theory

A curve showing frequency versus bias current for an increductor unit that is presently being produced is shown in Figure VIII-12. As is seen, a larger range of frequency than is required is available. This extra range can be used to increase the Q of the resultant circuit by enabling the use of a fixed inductor in series with the variable inductor

Since $Q = \omega L/R$, a high-Q inductor can be placed in series with the relatively low-Q variable inductor which will cause the inductances and resistances to add, but giving a proportion (or Q) which is higher than the original Q.
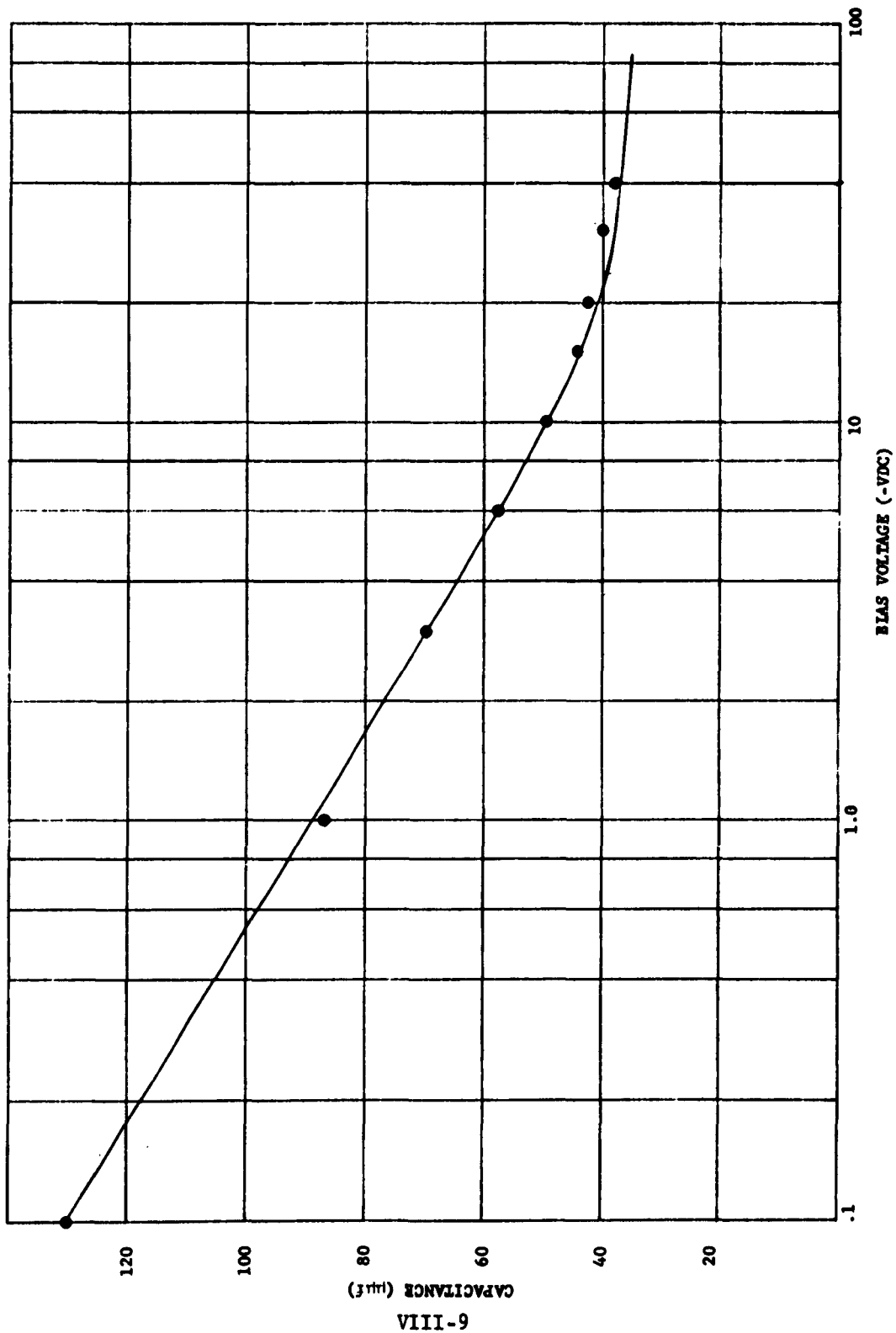
Figure VIII-6. Capacity vs. Bias Voltage Varicap #1
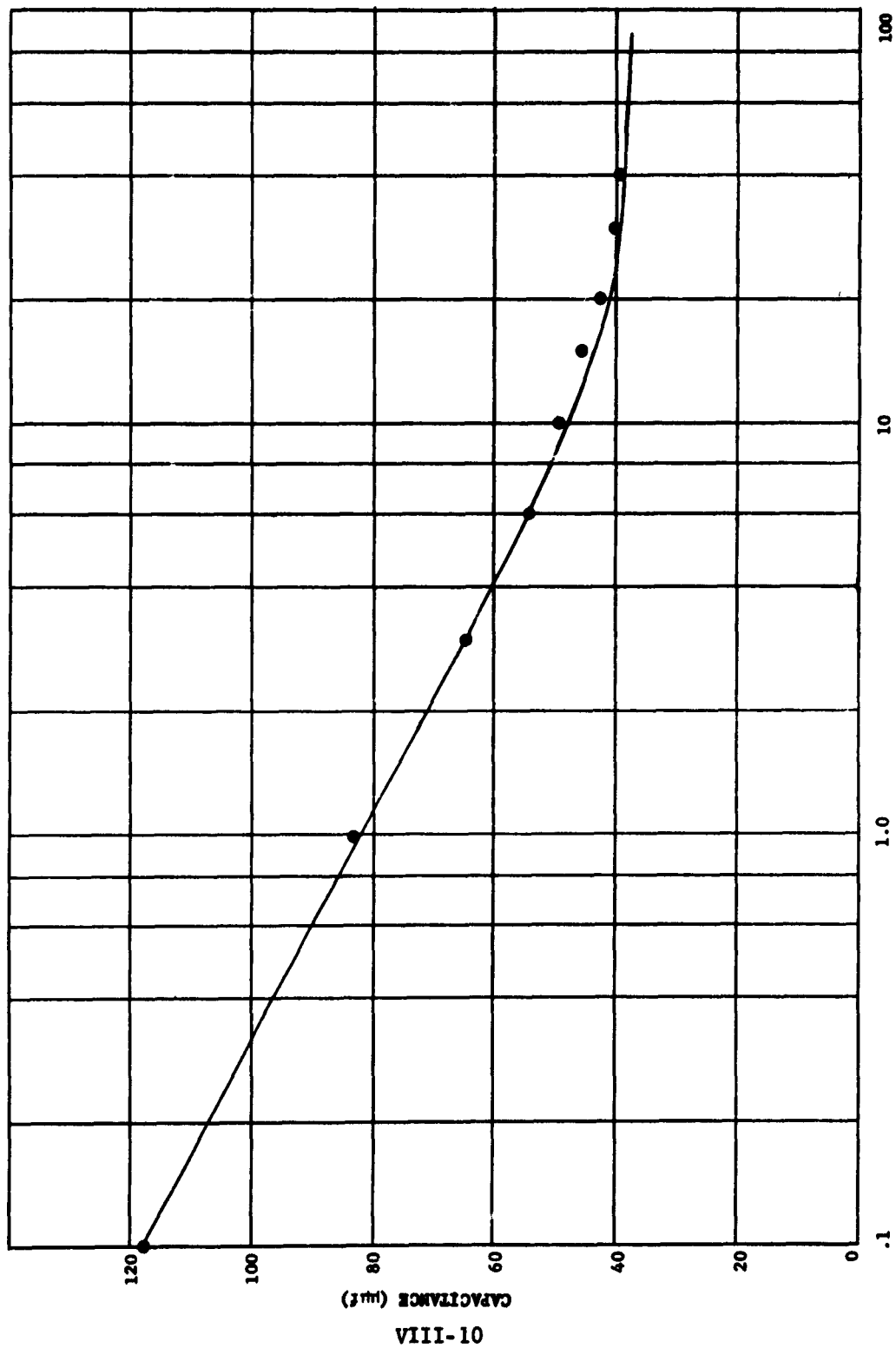
VIII-9

Figure VIII-7. Capacity vs. Bias Voltage Varicap #2

BIAS VOLTAGE (-VDC)

CAPACITANCE ($\mu\mu f$)
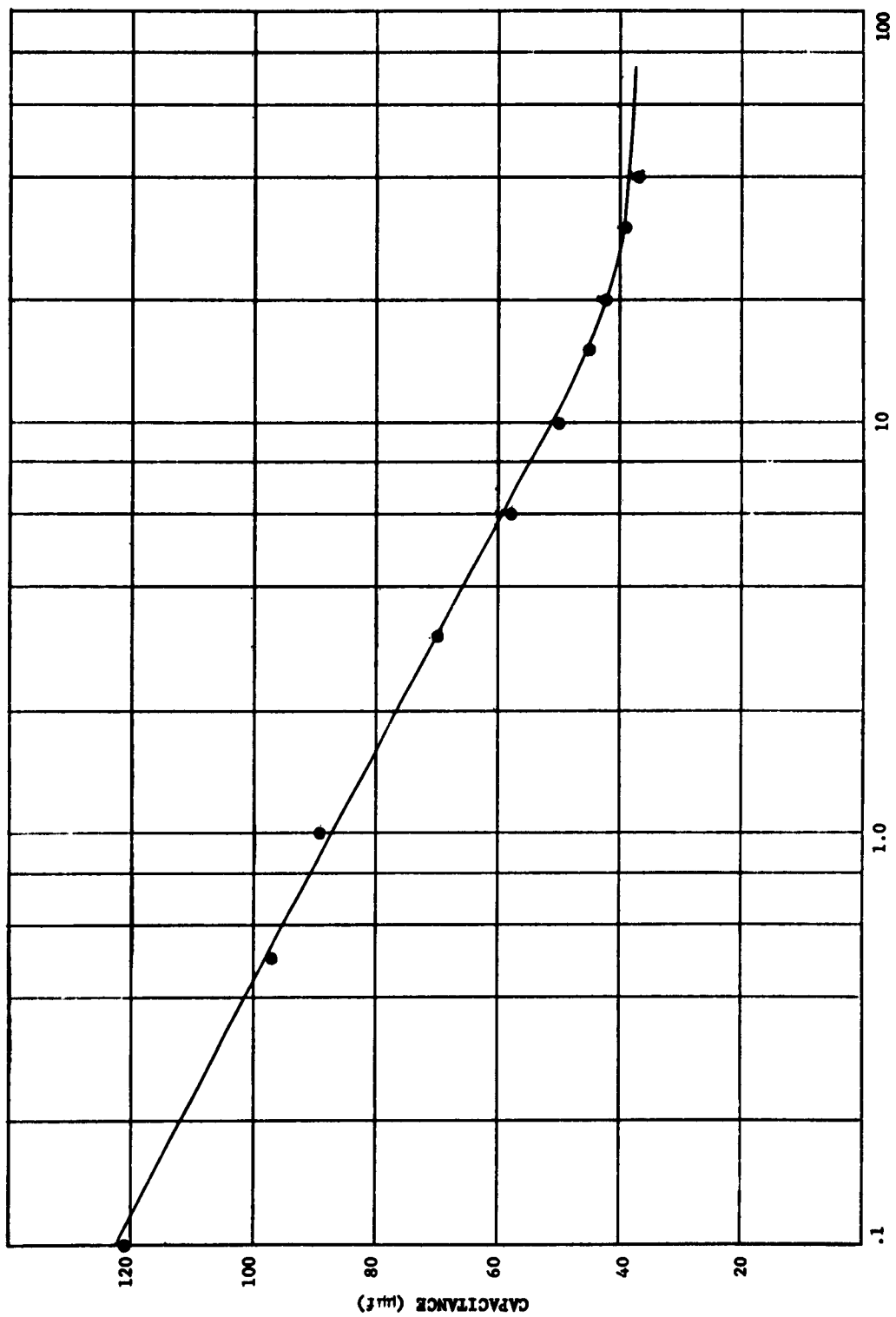
VIII-10

Figure VIII-8. Capacity vs. Bias Voltage Varicap #3
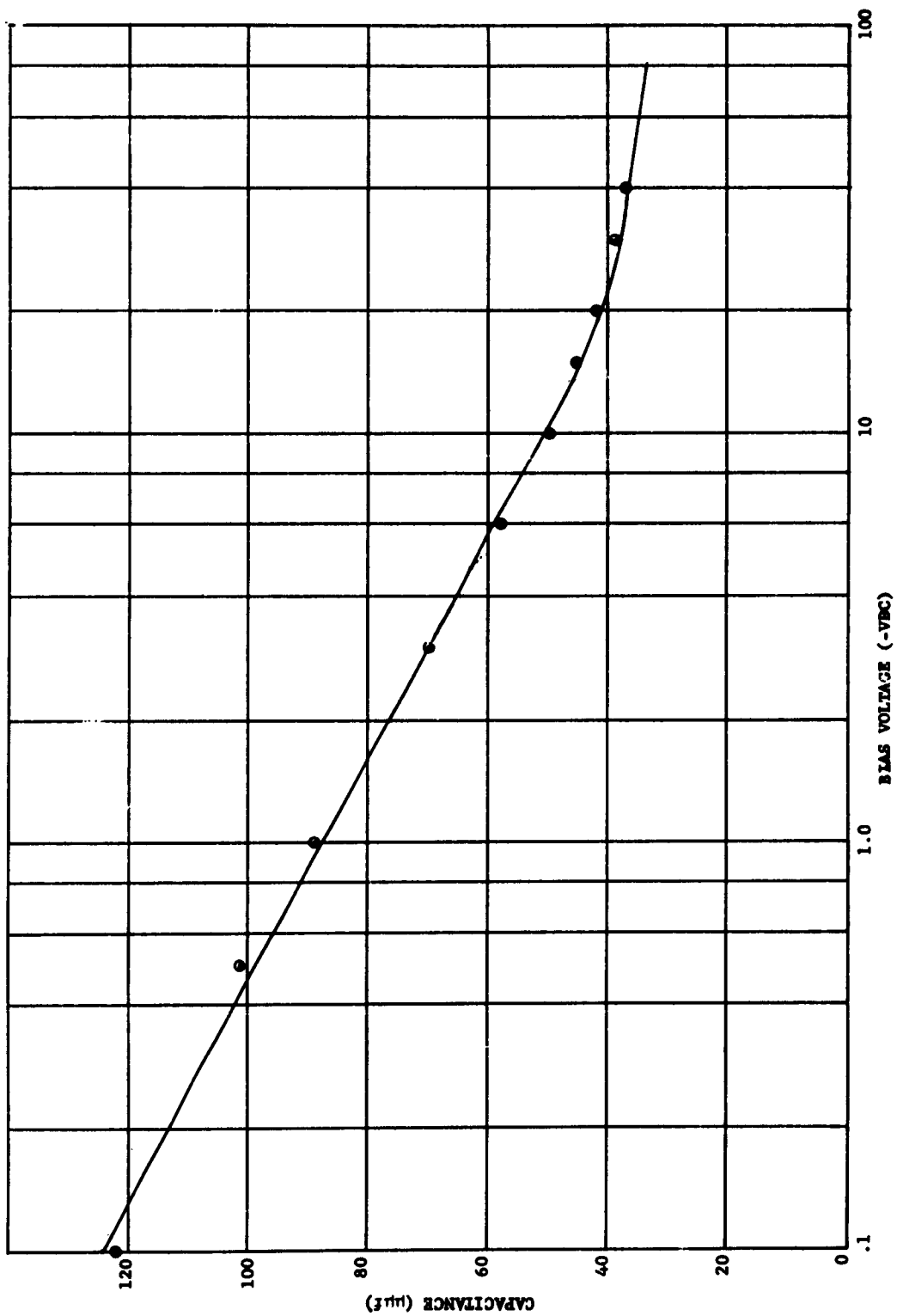
VIII-11

Figure VIII-9. Capacity vs. Bias Voltage Varicap #4
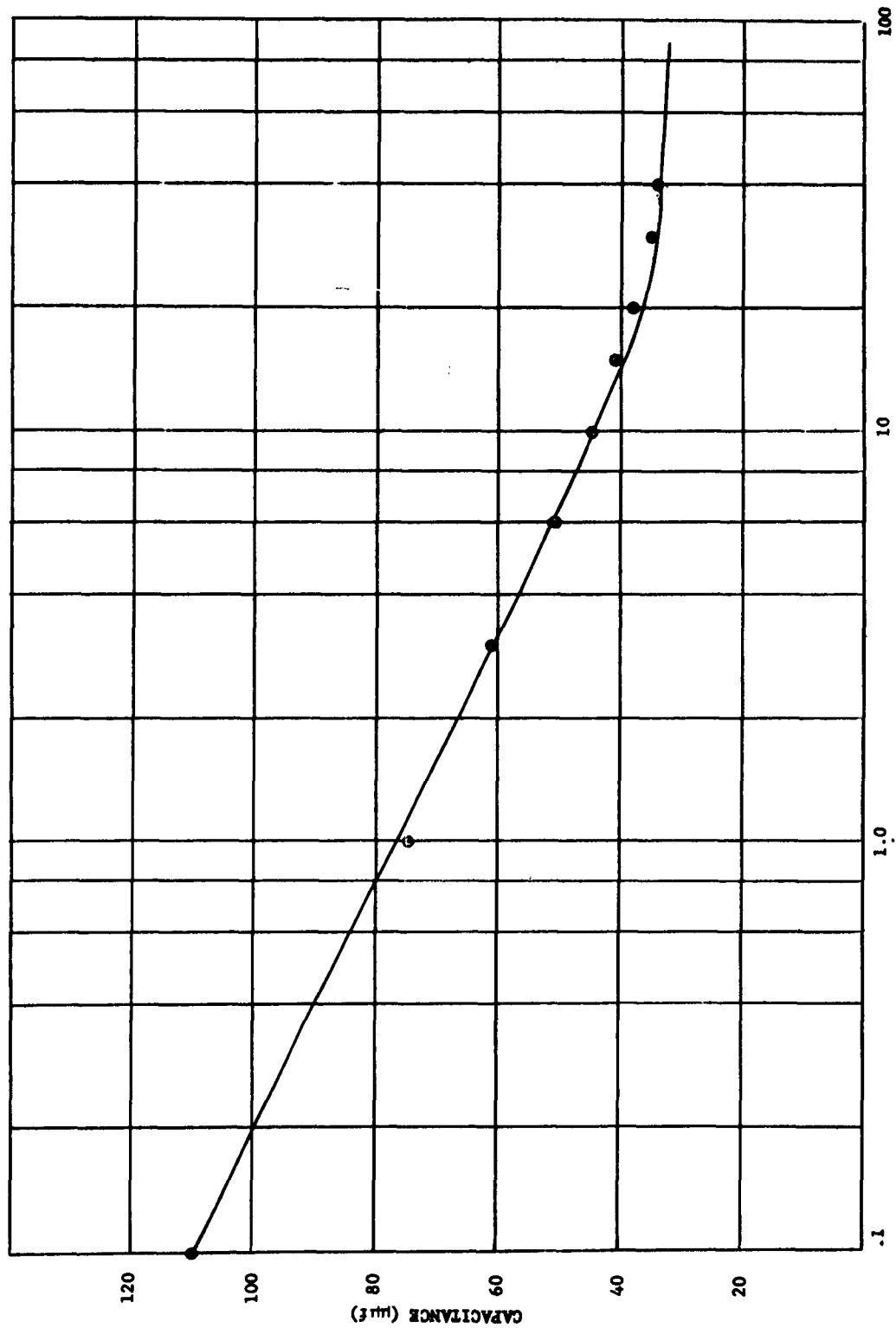
VIII-12

Figure VIII-10. Capacity vs. Bias Voltage Varicap #5
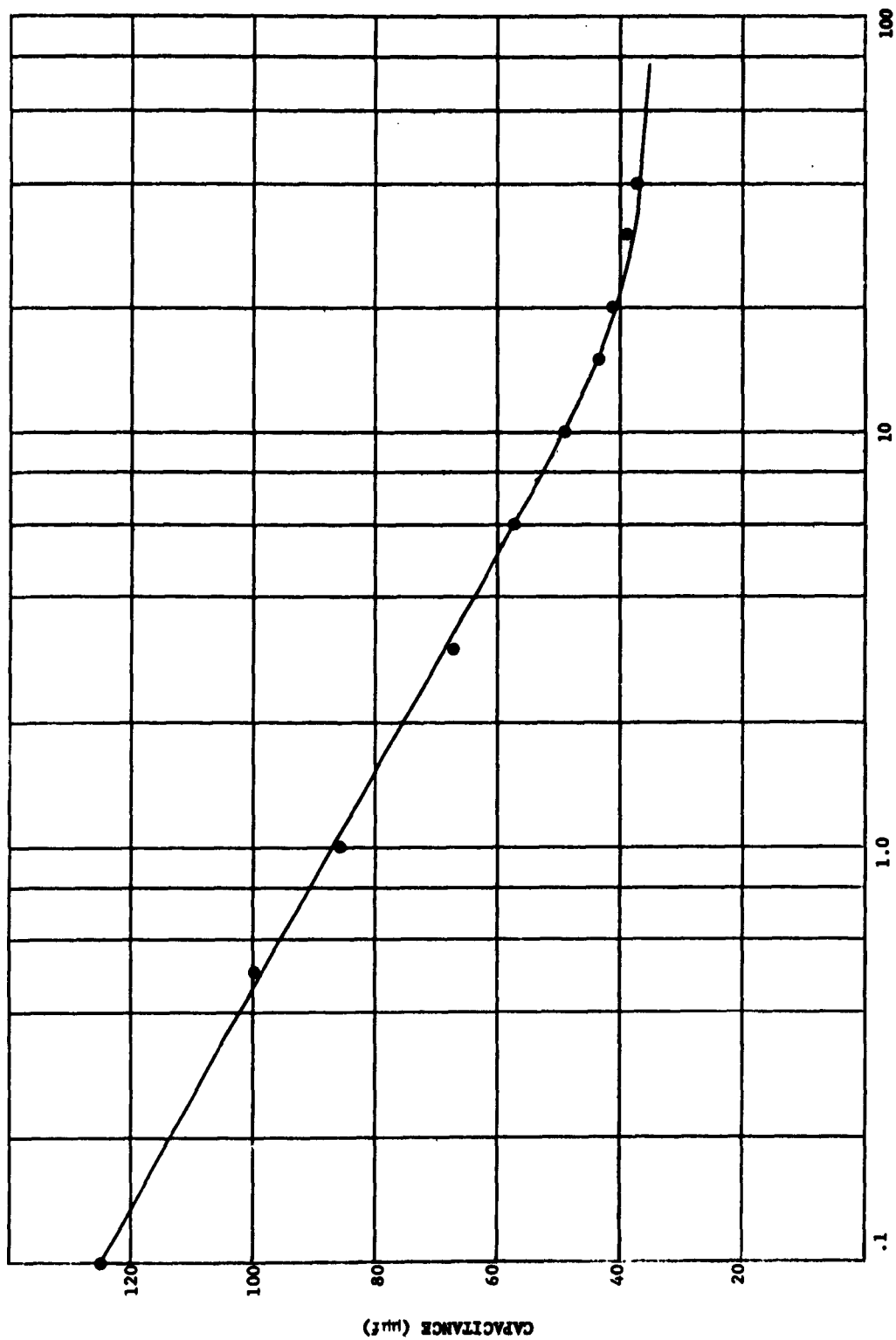
Figure VIII-11. Capacity vs. Bias Voltage Varicap #6
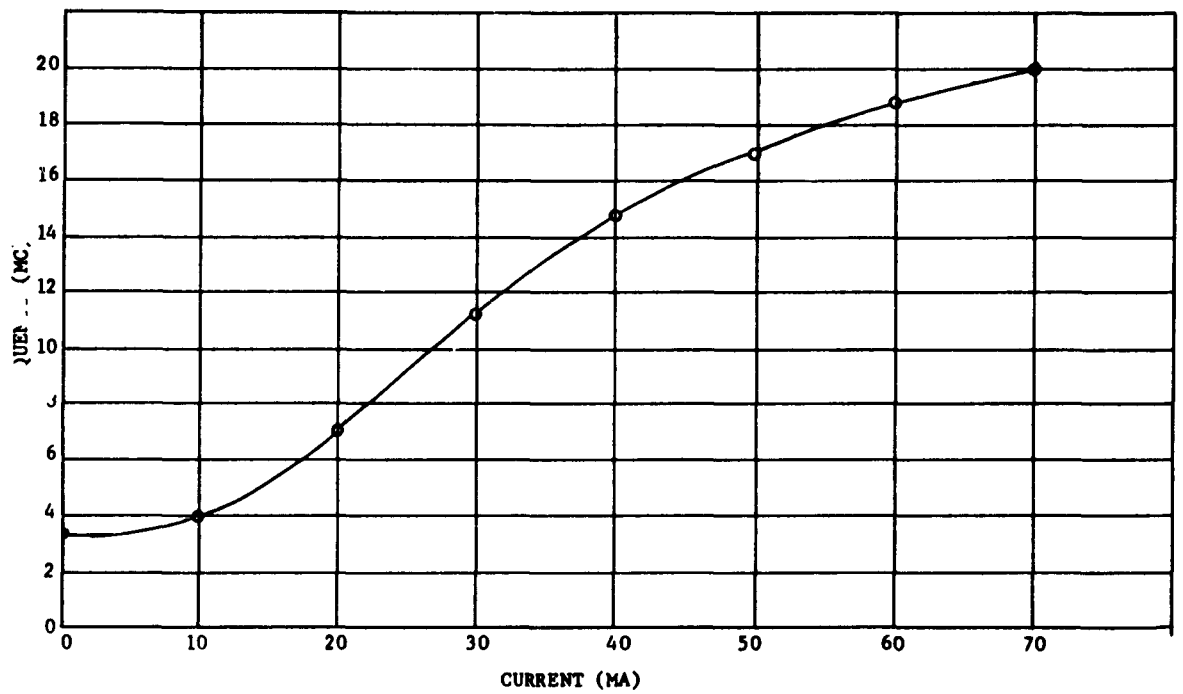
VIII-14

Figure VIII-12.   Frequency Versus Bias Current for an Increductor

VIII-15

Figure VIII-13, shows the variation in Q which is available from the same increductor unit which is referenced in Figure VIII-12. If care is taken with the circuit in which this variable inductor will be utilized, a circuit Q of 50 should be obtained.

As in the case of the oscillator, a bias of 0 to 70 ma will be needed for these increductor-tuned preselectors.

A feasibility breadboard has been built in the 8 Mc - 16 Mc band. Four sections of preselection, a mixer and a local oscillator are included in the breadboard shown in Figure 5 of the text. A circuit diagram is shown in Figure VIII-14.

Because of the construction of the "increductor" three-point tracking can be achieved if the proper circuitry accompanies the device. Besides the signal winding, a bias winding and control winding are built into the component. The bias winding is used to set the low frequency point while a trimmer capacitor can be used to set the high frequency. The slope of the frequency vs. bias current characteristic can be adjusted by means of a potentiometer in parallel with each control winding, thereby setting the mid-frequency range. Since the "increductor" is a current controlled device, the potentiometer in parallel with its control winding is acting as a by-pass for some of the control current. In order to insure that the low frequency end is held stable, a constant current source circuit was designed.
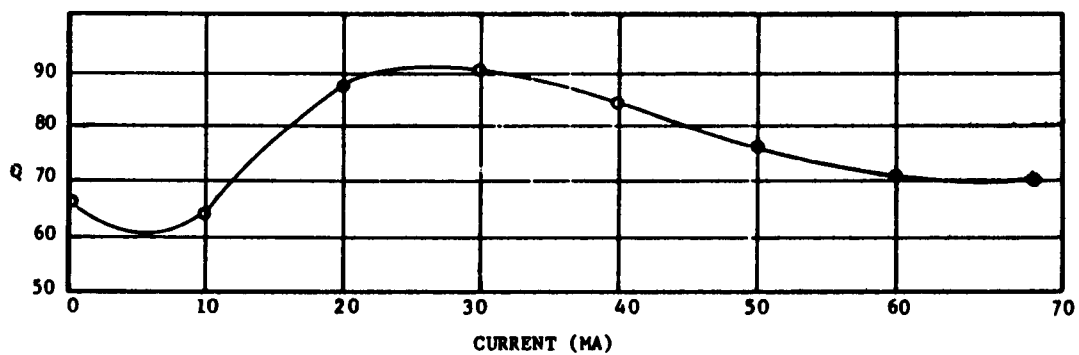
Figure VIII-13. Q Versus Bias Current for an Increductor
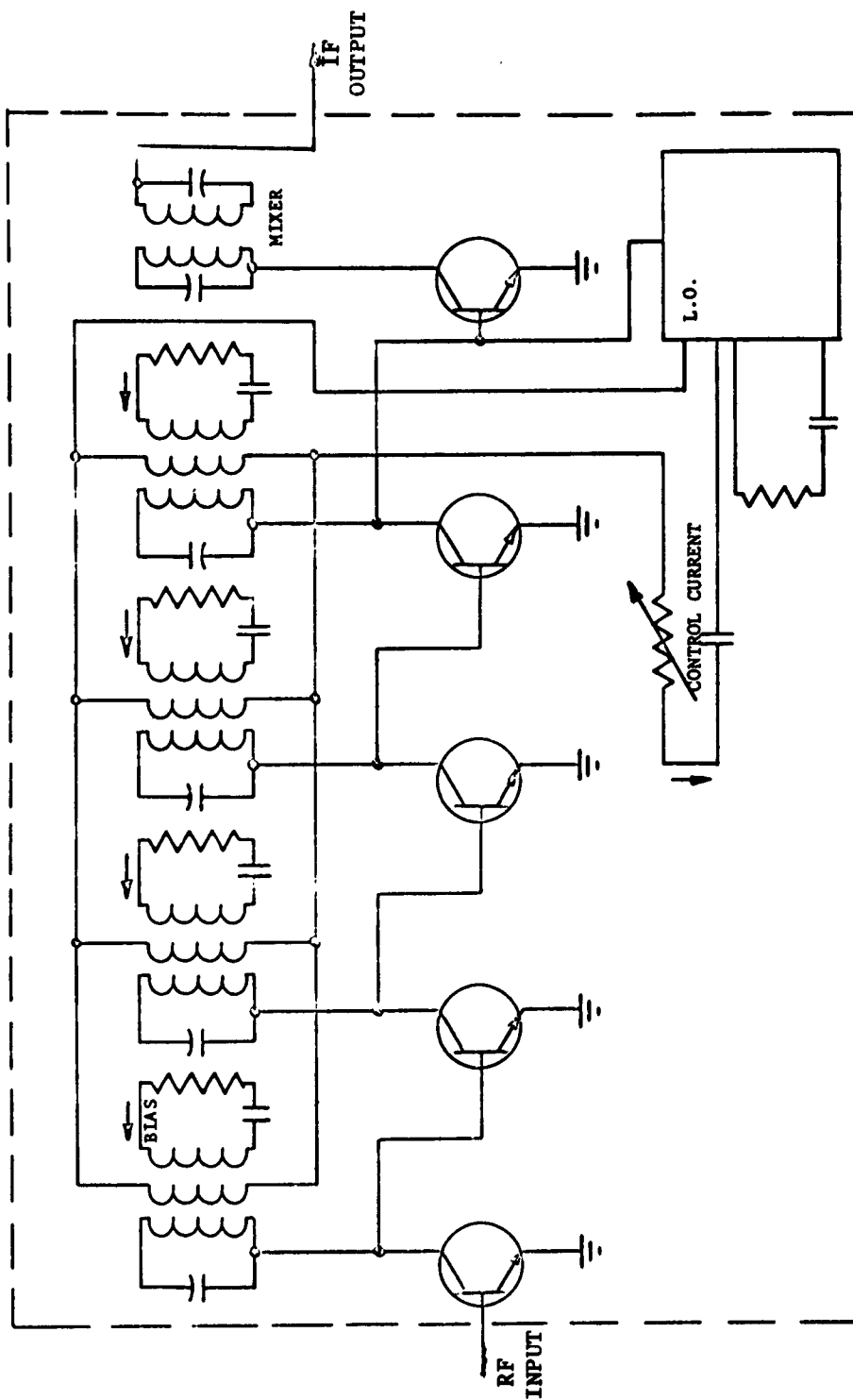
Figure VIII-14. Block Diagram, Increductor - Tuned Front End

VIII-18

APPENDIX IX

RF FRONT-END: 250 MC - 40 GC

A. GENERAL

The receiver RF front-end, as shown in Figure 2, consists of the RF preselector, the RF preamplifier, the local oscillator-converter, and the IF amplifier. The IF amplifier is not considered in detail in this report.

B. RF PRESELECTOR

As shown in Appendix III, the major technique for the elimination of spurious responses after the IF has been chosen is to use preselection. Since the receiver covers the frequency range of 100 kc to 40 Gc in twenty bands, it is desirable to utilize two types of preselection: fixed-tuned broadband band-pass filters and narrow-band tunable filters.

Band-pass filters normally have passbands at the 2nd or 3rd harmonic of the fundamental, which for the present application is extremely undesirable due to the large frequency range of the receiver. Increased harmonic-response-free range can be obtained by the use of complementary pairs of high- and low-pass filters. Harmonic frequencies to the 7th or 8th order are suppressed approximately 60 db with this technique. An off-band attenuation of greater than 60 db can be achieved at a frequency which is ± 5% away from the passband.

The response for a normalized, maximally flat Butterworth filter is

$$A = \sqrt{1 - X^{2n}} \qquad\qquad (IX-1)$$

where:

$$X = \frac{2(f-fo)}{BW}$$

and:

fo = center frequency

BW = 3-db bandwidth

n = number of poles in the filter's transfer equation.

A plot of this curve for $n = 1$ is shown in Figure IX-1. For an approximate plot, the asymptotes of this equation are plotted in Figure IX-2. For $X \gg 1$, the error in this approximation becomes negligible. The asymptotes are calculated as follows:

For $X \gg 1$

$$A(DB) = 20 \log (X^{2n})^{1/2}$$

$$= 20 \log X^{n}$$

$$= 20^{n} \log X$$

IX-1
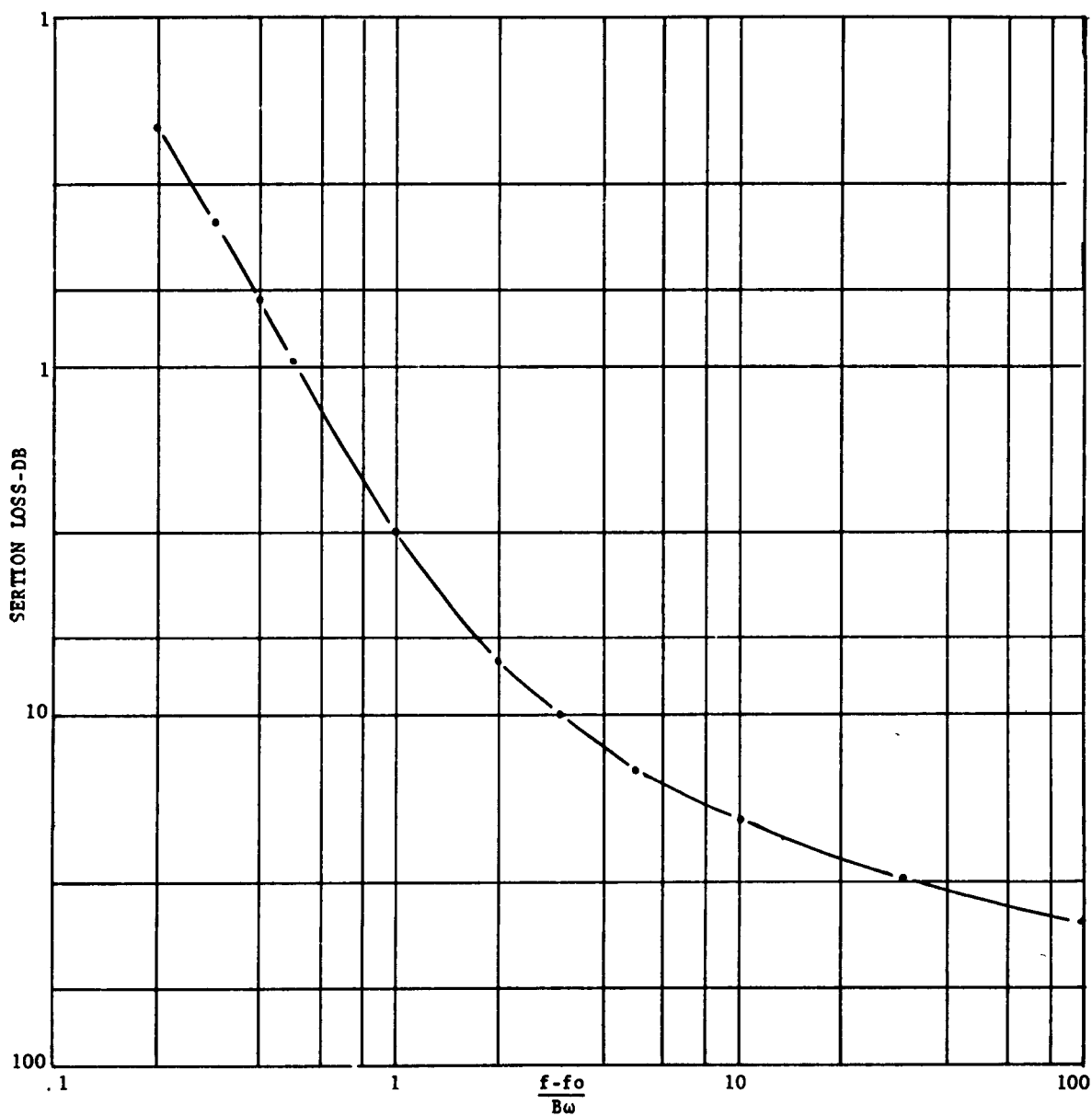
Figure IX-1. Insertion Loss vs $\frac{f-fo}{Bw}$ for N = 1

IX-2
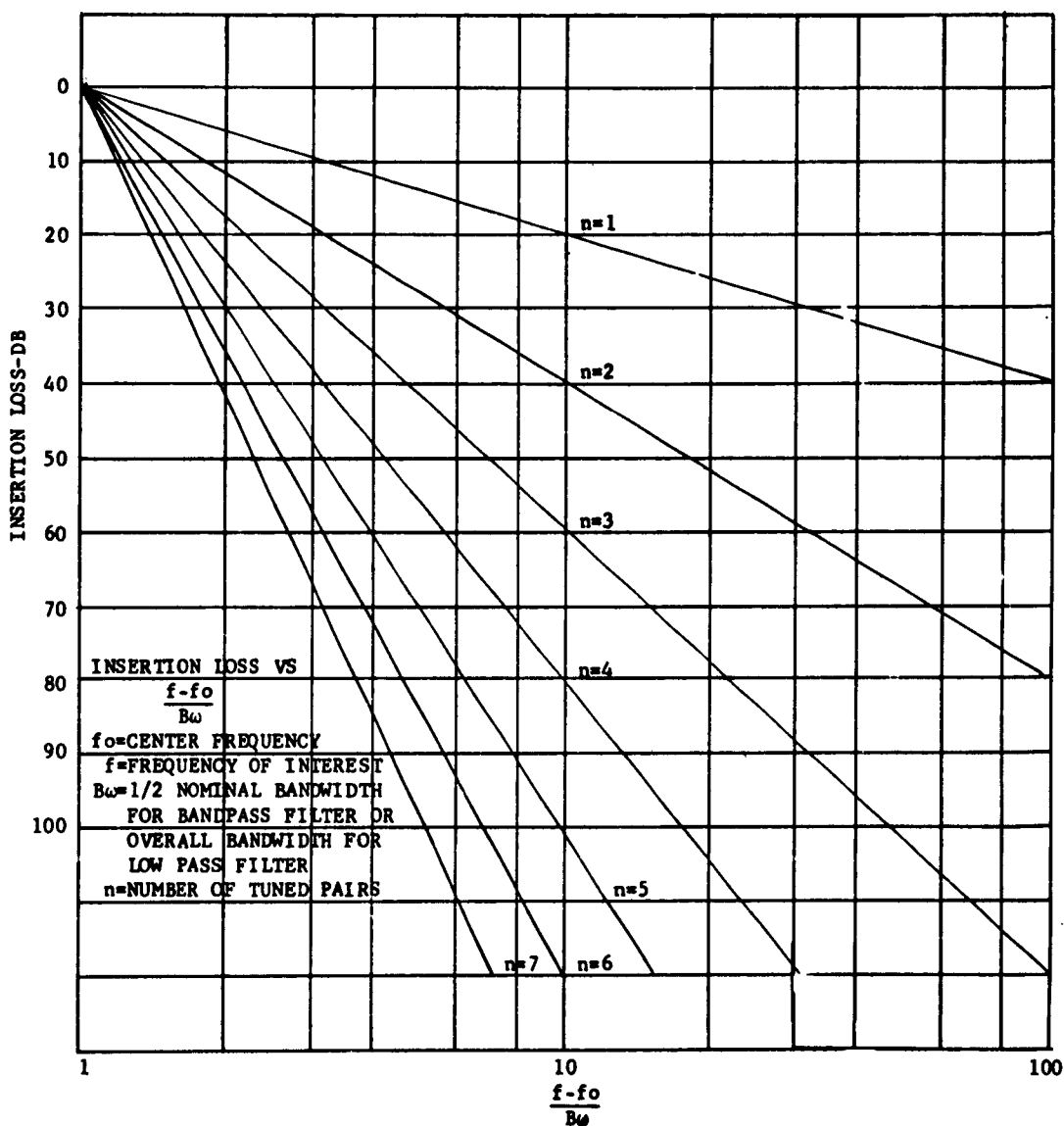
Figure IX-2. Plot of Asymptotes for $X = \dfrac{2(f-fo)}{BW}$

which says that the slope is 20 db per decade of bandwidth per tuned pair. If sharper skirts are needed, M-derived filters can be used, but the flatness of the filter characteristic will be sacrificed.

Appendix VI showed that the local oscillator, and, therefore, the preselector, must be swept at a speed which is beyond the capability of a mechanically swept system. Therefore, the preselector must be capable of being electronically swept over an octave bandwidth as well as tracked with the receiver local oscillator.

For the lower frequencies (up to approximately 400 Mc) electrically controlled preselectors employing varicaps or increductors will be utilized. Varicap-controlled preselection is described in Appendix VIII. Increductor controlled preselectors are now being investigated and will be reported on later.

For frequencies above 400 Mc, a Yttrium-Iron-Garnet (YIG) or Lithium-Ferrite preselector will be utilized. For the past several years, considerable research effort has been spent on these devices at Watkins-Johnson, Hughes, Stanford Research, Sperry, Melabs, Loral, etc. Lithium-Ferrite preselectors have been built for frequencies between 160 Mc and 2 Gc, and YIG preselectors are commercially available for frequencies between 2 Gc and 85 Gc.

The ferrite crystal type of preselector utilizes the unpaired electron spin in the ferrite material to cause coupling of RF energy through a cross-coupled transmission line structure; crossed coaxial or strip transmission lines or loops, and waveguide-beyond-cutoff. When a uniform DC magnetic field is applied to a ferrite crystal in the Z direction, and an RF magnetic field in the X direction, a gyroscopic effect is produced, causing the magnetic moment vector of the electron to precess about the Z axis. Such electron precession causes the RF energy to be coupled out in the Y direction. By changing the DC field, the frequency of coupling may be changed.

From the insertion loss formula for a single tuned filter,

$$\text{I.L.} = \frac{1}{(1 - Q_\ell/Q_u)^2} \quad , \qquad \text{where } Q_\ell = \text{loaded Q} \qquad \text{(IX-2)}$$

$$Q_u = \text{unloaded Q} = \frac{fo}{\Delta f}$$

the insertion loss of a ferrite-coupled filter may be seen to be

$$\text{I.L.} = \frac{1}{(1 - \frac{\Delta f}{B})^2} \approx 1 + 2\left(\frac{\Delta f}{B}\right) \text{ when } \frac{\Delta f}{B} \ll 1 \qquad \text{(IX-3)}$$

This simple expression indicates the dependence of insertion loss on bandwidth. As bandwidth increases, the insertion loss decreases. Figure IX-3 illustrates the relation between insertion loss and bandwidth that one can expect with actual single-tuned YIG filters between 2 Gc and 10 Gc.
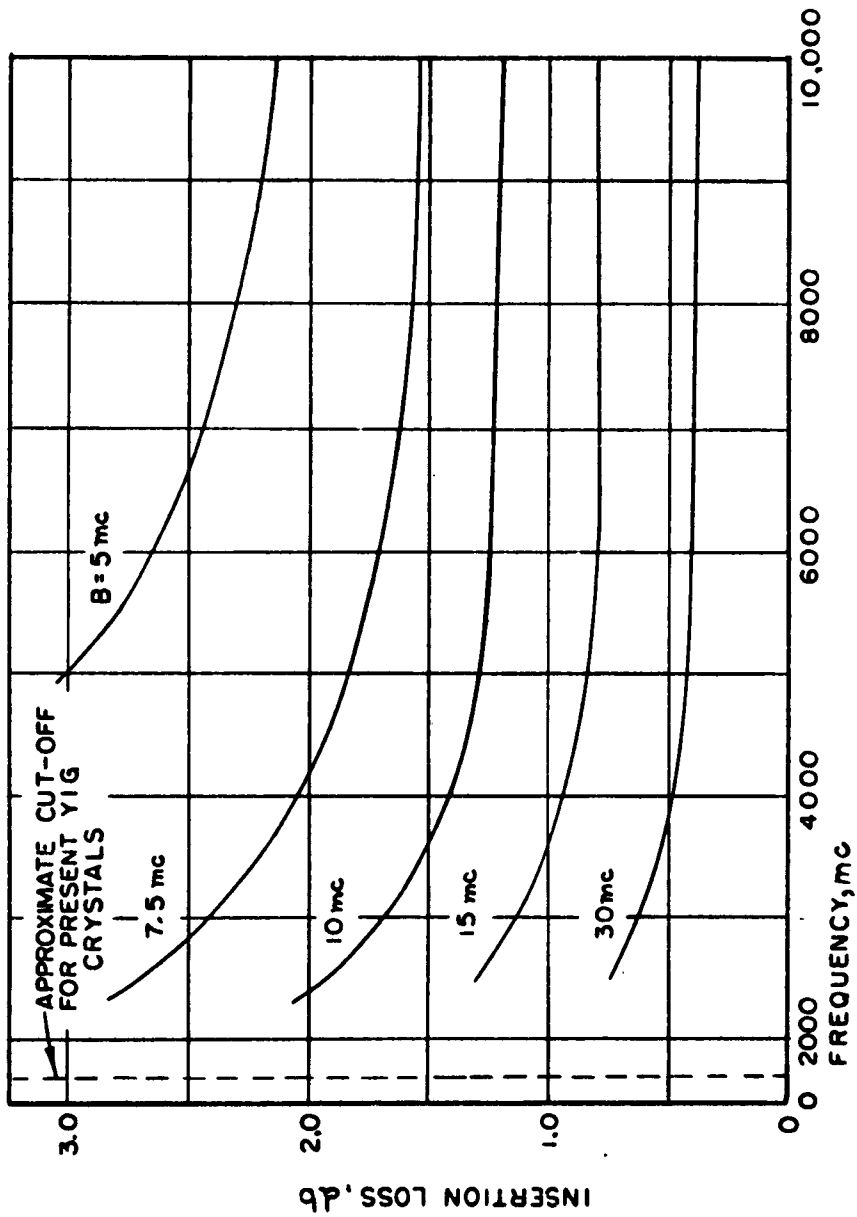
Figure IX-3. Insertion Loss of a Single-Tuned YIG Filter as a Function of Frequency and Bandwidth

1101436

Much of the present research is in the materials area in an effort to extend the usable frequency range and to minimize insertion loss. For the present 2-Gc to 40-Gc application, YIG crystals appear to offer the optimum preselector properties.

It has been shown previously, in Appendix V, that a four-section preselector is required in order to obtain the desired image and local-oscillator radiation suppression. Quadruple-tuned YIG filters built by Watkins-Johnson utilize the principle of waveguide-below-cutoff to achieve off-resonance isolation. The input and output coaxial lines are connected by a short section of guide which is well beyond cutoff for the desired operational band. The YIG crystals are placed in this cutoff waveguide and the degree of coupling is controlled by the size and spacing of the crystals. A possible configuration is shown in Figure IX-4.

Watkins-Johnson has built and tested several four-section pre-selector filters above 2 Gc. Typical specifications for a K-band unit are:

Frequency range	12-18 Gc
Insertion loss	< 3 db
Bandwidth (3 db)	30 Mc
Skirt selectivity	24 db/octave
Off-resonance isolation	> 50 db
Spurious level	- 30 db or better
Directivity	> 20 db
DC power for 3 Gc tuning	10 watts
Size	2.1" x 2.6" x 4"
Weight	3 pounds

Tracking between the four filter sections poses no problems since the four crystals are tuned via a common coil. Tracking between the local oscillator and the preselector can be accomplished across an octave bandwidth with less than a 0.1 percent deviation of filter center-frequency from the sum of the local oscillator plus the inter-mediate frequency. The tuning curve of current vs. frequency for the filter is extremely linear, and that of the tunable local oscillator can be made linear by using a linearizing network, if necessary. Present state-of-the-art will allow the sweeping of an octave bandwidth at rates above 1000 cps.

The YIG filter exhibits two extremely useful characteristics: non-reciprocity and frequency-selective limiting.

Non-reciprocal YIG filters act as isolators. This effect has allowed as many as four filter sections to be built in a single unit.
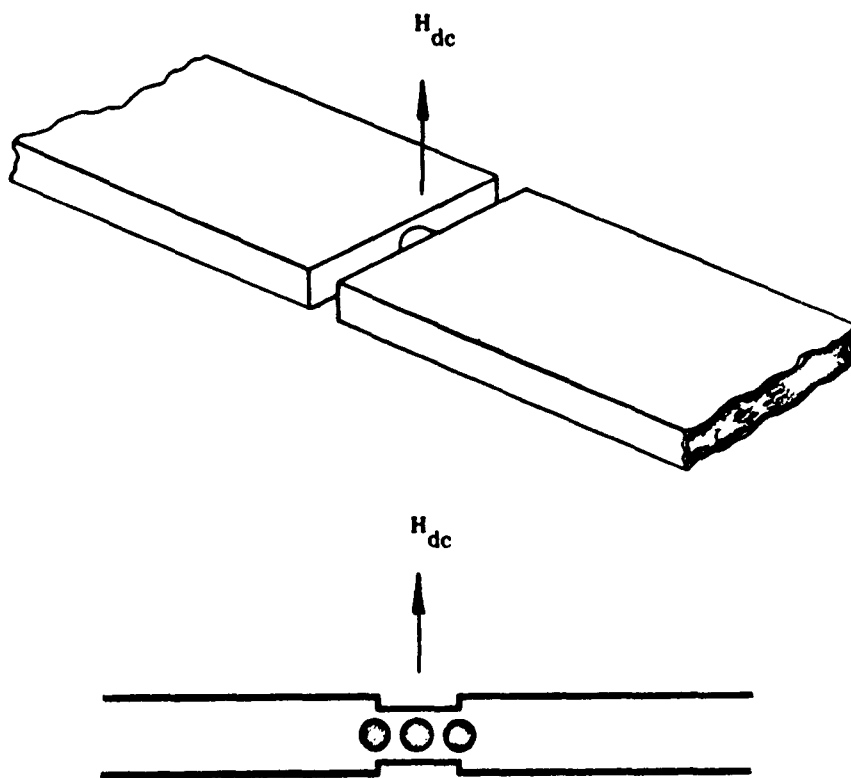
Figure IX-4. Possible Circuit Configuration for a
Multiple-Tuned YIG Bandpass Filter in Waveguide

Moreover, the isolation produced by the YIG filter will aid in the reduction of spurious signals which are transmitted from a receiver; for example, the reduction of local-oscillator radiation.

Frequency-selective limiting is a very useful property for incorporation in the front-end of a broadband receiver. This property, which stems from the large number of closely spaced spin modes inherent in a YIG preselector, allows the limiting of a strong signal and the simultaneous reception of a weak signal displaced by only a few megacycles. No additonal attenuation is presented to the weak signal in the presence of the strong limited signal.

Several minor problems are associated with the presently available YIG filters.

YIG filters will change center frequency on the order of the bandwidth (25 Mc) with a temperature change of -20°F to +132°F. This change can be held to less than 5.0 Mc by maintaining the temperature of the YIG crystal above 20°C. Still further reduction can be obtained by holding the YIG at higher temperatures. The Curie temperature of a Yttrium-Iron-Garnet mixture is 275°C.

The hysteresis effect inherent in the YIG tuning structure will cause the center frequency to shift when the filter is tuned to a given frequency from opposite directions. This shaft will be less than 20 Mc and can be eliminated by shaping the driving-voltage waveform to return the filter to the same point on the hysteresis curve at the start of each sweep cycle.

Perhaps the most serious limitation of the YIG devices is the spurious responses. Spurious responses on the order of -30 db down will exist in the octave bandwidth. The position of the spurious response (usually only one) bears a fixed relationship to the center frequency. This fact, coupled with the IF selectivity, minimize the effect of spurious responses in the preselector. Moreover, utilization of an RF preamplifier between two filters will double the db rejection of a single filter.

C. RF PREAMPLIFIER

1. Tube Amplifiers

Distributed-parameter tube-type amplifiers and traveling-wave tube amplifiers are state-of-the-art devices whose characteristics are well known. No effort has been devoted to the study of these devices for the present program.

2. Transistor Amplifiers

Low-noise, transistorized amplifiers with better than octave bandwidths are commercially available up to 1000 Mc. These amplifiers are ideal for solid-state amplification as shown by the following characteristics:

Noise Figure: 4 db at 100 Mc to 10 db at 1000 Mc (Lower noise figures can be obtained with amplifiers of narrower bandwidths.)

Bandwidth: Octave or better

Gain: 30 db or better

3. Tunnel-Diode Amplifiers

Tunnel-diode amplifiers show potential for the entire frequency range up to 40 Gc. During this quarter of the program, the tunnel-diode amplifier investigation has been limited to an analysis of the amplifier circuit.

The following paragraphs present a treatment of the basic amplifier design approach recommended by AEL.

a. Amplifier Analysis

The following is a treatment of the basic amplifier design procedure employed in the tunnel-diode amplifier investigation at AEL.

Figure IX-5 represents an idealized i-e characteristic for the tunnel diode. From this characteristic the negative conductance is calculated to be:

$$g = \frac{2(I_v - I_p)}{V_v - V_p} \quad \text{mhos} \tag{IX-4}$$

It will be seen from Figure IX-5 that the conductance varies continuously in the negative region so that Equation IX-4 is only a good approximation to "g" at the point of maximum slope. By having an arrangement for varying the diode bias, the conductance, as calculated by Equation IX-4 can be found.

I_p = peak diode current

I_v = Valley current

V_v = Valley voltage

V_p = Peak current voltage intercept.
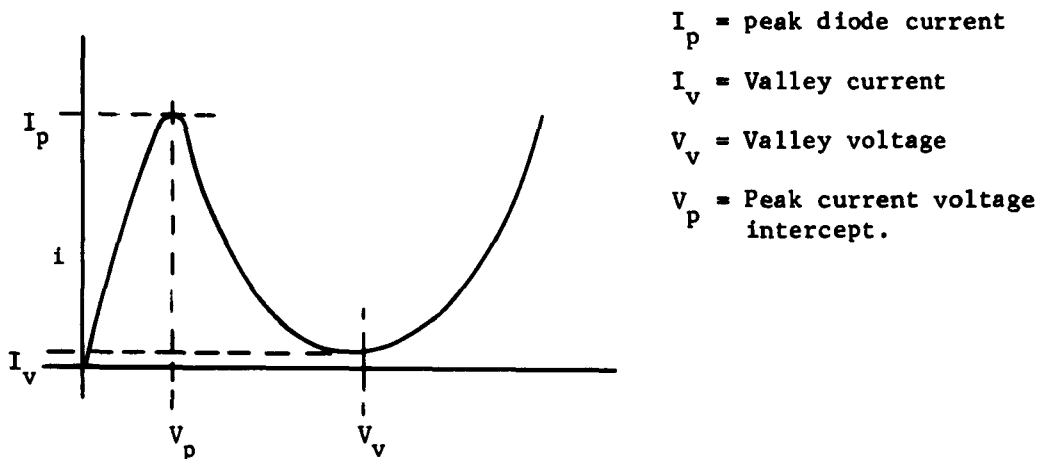
Figure IX-5. Typical Idealized Tunnel-Diode Characteristic

We now consider the tunnel diode as an active element in a single
loop circuit. Figure IX-6 indicates a typical circuit where the diode
is in series with the load and generator impedances. An analysis
which is similar to that which follows could be made with the tunnel
diode in parallel with the load and generator impedances.

$*L_S$ = Tunnel diode lead
inductance

R_G = Generator impedance

R_L = Load impedance

$*R_S$ = Spreading resistance

L_C = External circuit
inductance

$*C$ = Barrier capacitance

$*-g$ = Barrier conductance

$L = L_S + L_C$

$R_T = R_L + R_G + R_S$

a) Diode and loop equiva-
lent circuit

b) Equivalent circuit with
all resistance and induc-
tance lumped.
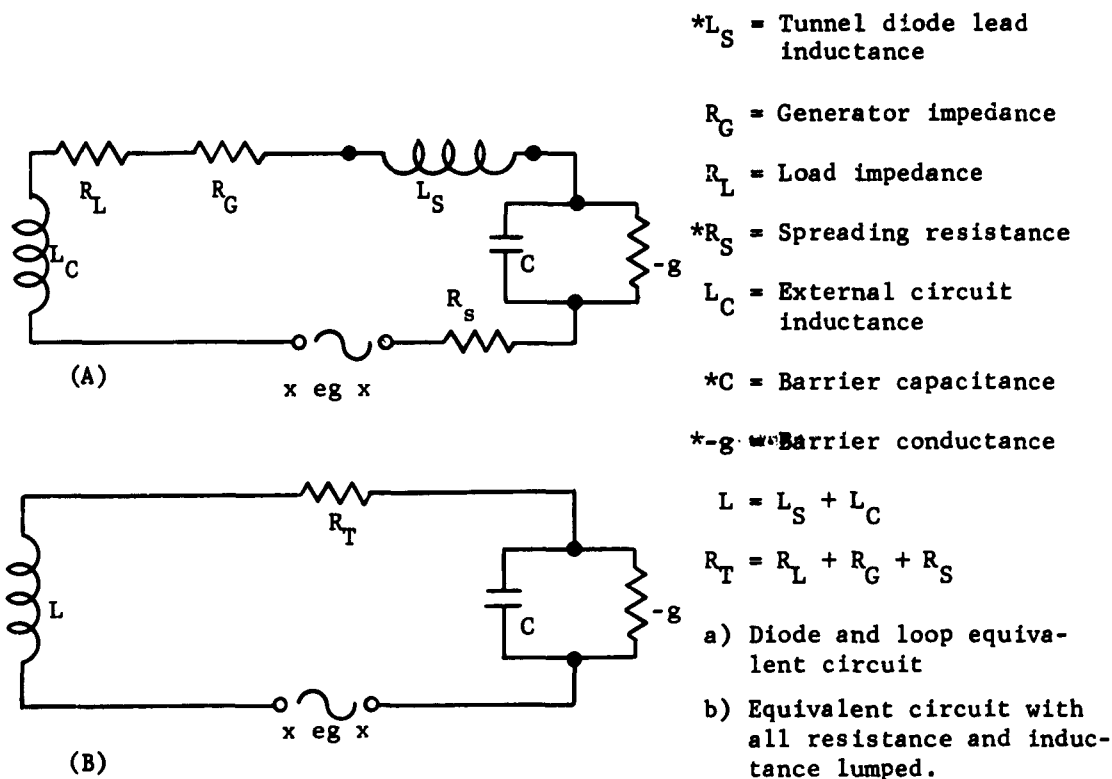
Figure IX-6. Tunnel-Diode Equivalent Circuit

In Figure IX-6, two assumptions have been made for simplicity:

(1) It is assumed that the generator and load impedances
are real;

(2) It is further assumed that the external circuit reactance
is positive.

All series reactance must be cancelled for maximum gain, as will be seen.
Any reactance appearing in the load, generator, and external circuit
can be combined into a single reactance for analytic purposes. This
combined reactance will be positive.

The Nyquist criteria for amplifier stability requires the zeroes
of the circuit input characteristic impedance (terminals x-x in Figure
IX-6b) must fall in the left half of the complex frequency plane.

Therefore:

$$Z_i(S) = \frac{S^2 LC + (R_T C - Lg)S + (1 - R_T g)}{SC - g} \quad \text{ohms} \tag{IX-5}$$

The zeroes of (2) are:

$$S = \frac{1}{2}\left(\frac{R_T}{L} - \frac{g}{c}\right) \pm \sqrt{\frac{1}{4}\left(\frac{R_T}{L} - \frac{g}{C}\right)^2 - \frac{1 - R_T g}{LC}} \tag{IX-6}$$

Clearly, if a zero is to appear in the left half plane, the following must be true:

$$\frac{R_T}{L} - \frac{g}{C} > 0 \tag{IX-7}$$

and

$$1 - R_T g > 0 \tag{IX-8}$$

and

$$\frac{Lg}{C} < R_T < \frac{1}{g} \tag{IX-9}$$

Expressions (IX-7), (IX-8), and their combination, expression (IX-9), represent the stability requirements for the tunnel-diode amplifier.

For completeness, the circuit in Figure IX-6b will oscillate if a zero in expression (IX-6) lies on the real-frequency axis of the complex-impedance plane. Therefore, for oscillation,

$$S = \sigma + j\omega \tag{IX-10}$$

$$\sigma = 0$$

$$\frac{R_T}{L} - \frac{g}{C} = 0$$

The frequency of oscillation will be given by:

$$\omega = \frac{1 - R_T g}{LC} \quad \text{radians/sec} \tag{IX-11}$$

The appropriate Z_i -vs.-ω plots appear in Figure IX-7 for the stable amplifier and the oscillator.

Substituting $j\omega$ for S in expression (IX-5) yields the input impedance for the circuit in Figure IX-6b.

$$Z_i = (R_T - \frac{g}{\omega^2 C^2 + g^2}) + j\omega (L - \frac{C}{\omega^2 C^2 + g^2}) \text{ ohms} \qquad \text{(IX-12)}$$

(a) STABLE AMPLIFIER (b) FREE OSCILLATOR
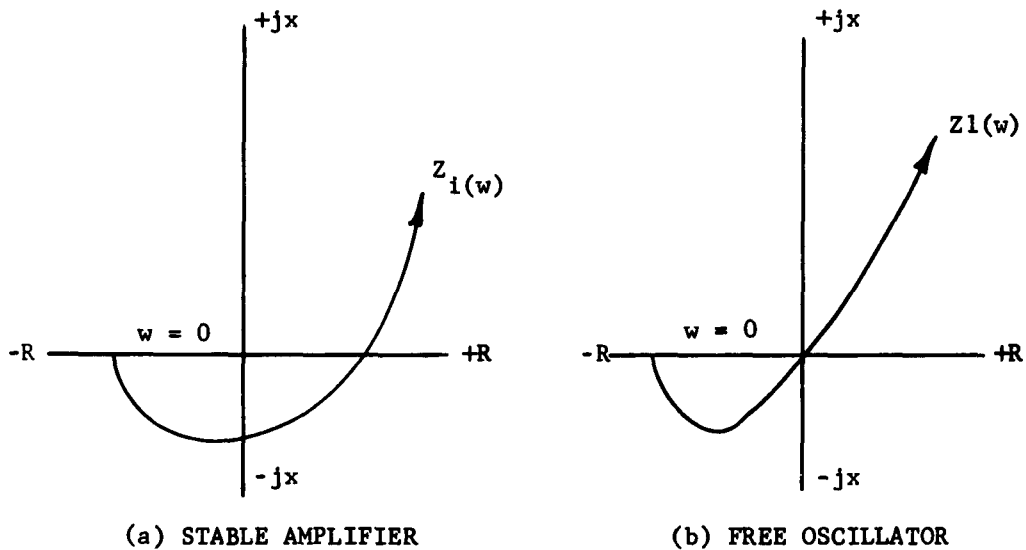
Figure IX-7. Complex Plane Impedance vs. Angular Frequency
for (a) Stable Amplifier, (b) Free Oscillator.

The insertion gain (G) of a tunnel-diode amplifier is defined as:

$$G = \frac{P_d}{P_o} \qquad \qquad \text{(IX-13)}$$

where: P_d = Power delivered to
the load with the
tunnel diode in
the circuit

P_o = Power delivered to
the load without
the tunnel diode
in the circuit.

$$P_d = eg^2 \frac{Z_L}{Z_i^2} \text{ watts} \qquad \text{(IX-14)}$$

$$P_o = eg^2 \; \frac{Z_L}{Z_o^2} \;\; \text{watts} \hspace{2cm} \text{(IX-15)}$$

where: Z_o = total circuit impedance without the tunnel diode

Finally, combining (IX-13), (IX-14), and (IX-15):

$$G = \frac{P_d}{P_o} = \left(\frac{Z_o}{Z_i} \right)^2 \hspace{2cm} \text{(IX-16)}$$

Gain will occur when the total circuit impedance is reduced by the addition of the tunnel diode.
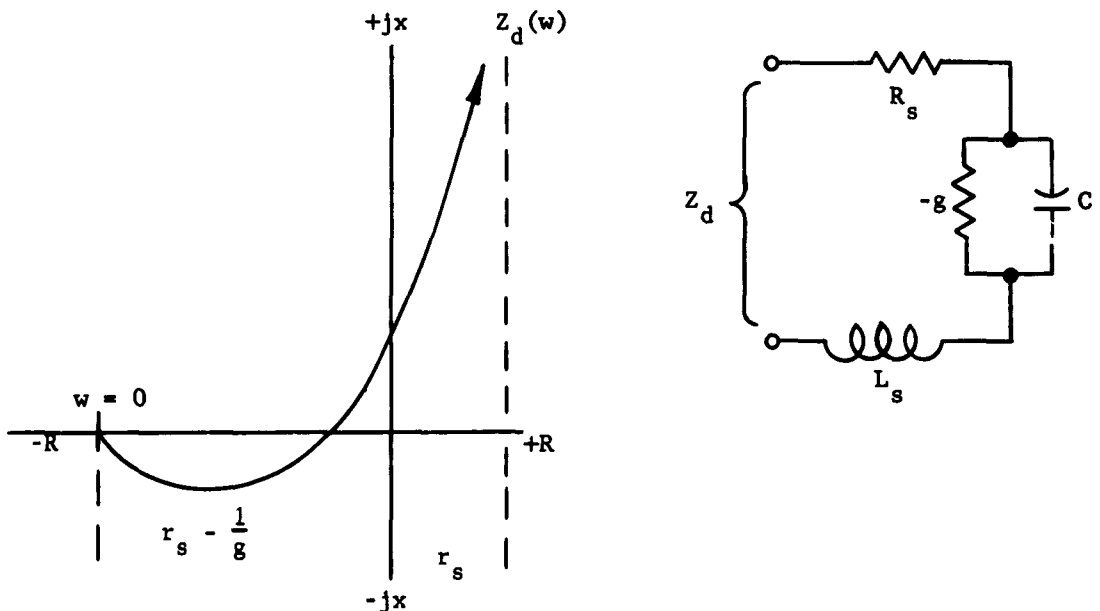
Figure IX-8. Typical Tunnel-Diode Input Impedance vs. Frequency

A graphic presentation of the tunnel-diode input impedance is shown in Figure IX-8. The diode alone is unstable and will self-oscillate at the point (frequency) where the $Z_d(\omega)$ curve crosses the real axis. Another point of interest is the point where the $Z_d(\omega)$ curve crosses the imaginary axis. At this point, the negative conductance of the diode is reduced to zero and no further amplification is possible. This is referred to as resistive cut-off.

The gain-bandwidth characteristic of tunnel-diode amplifiers is best shown by a graphic presentation of the tunnel diode input impedance as modified by the addition of the required circuit elements to achieve amplification. This is done in Figure IX-9.

Figure IX-9a again presents the input impedance versus frequency of the tunnel diode. From Figure IX-6b and expression (IX-9),

$$R_T = \frac{g}{\omega^2 C^2 + g^2} \quad \text{and } L = \frac{C}{\omega^2 C^2 + g^2} \qquad \text{(IX-17)}$$

We have the input impedance plot for the amplifier circuit as shown in Figure IX-9b. If a radius vector (b) is drawn from the origin to the $Z_i(\omega)$ curve, it will represent the amplifier input-circuit impedance. If a circle is now drawn representing the magnitude of the load and generator impedances at a radial distance (a) from the origin, the gain of the amplifier can be shown, as in Figure IX-9c, to be

$$G = \left[\frac{a}{b}\right]^2 = \left[\frac{Z_o}{Z_i}\right]^2$$

Notice that the points ω_1 and ω_2 represent the angular frequencies in Figure IX-9b at which gain is unity. At all frequencies between ω_1 and ω_2, the circuit amplifies.

If expression (IX-14) is satisfied, then there will be infinite gain at the selected ω_o. In practice, gain approaching infinity is is not attainable. Expression (IX-14), precisely satisfied, is the boundary between infinite gain and free oscillation. Figure IX-9b corresponds to Figure IX-7b.

A 2-Gc to 4-Gc tunnel-diode amplifier will be designed, following the above procedure, later in the program.
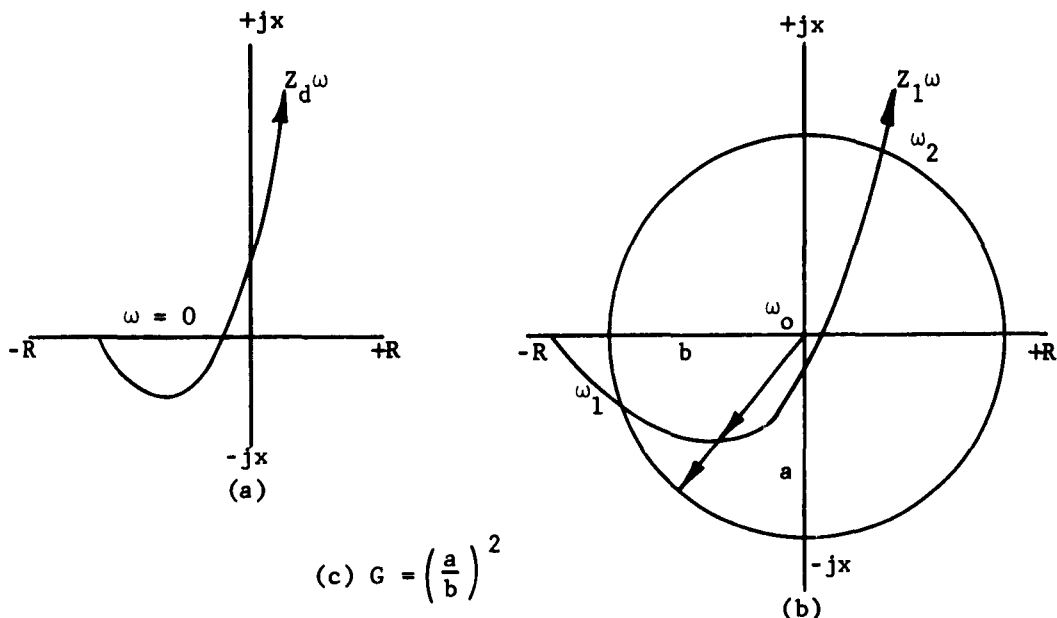
Figure IX-9. Gain-Bandwidth Characteristic of Tunnel-Diode Amplifiers

a) $Z_d(\omega)$

b) $Z_i(\omega)$ with external circuit elements adjusted to produce gain.

c) Gain calculation from b.

D. LOCAL OSCILLATOR-CONVERTER

1. Tube-Type Oscillator with Crystal Mixers

Backward-wave oscillators, voltage-tunable magnetrons, thermally-tuned klystrons, etc. are available over the entire 250-Mc to 40-Gc range. However, many systems have been built by using these oscillators and crystal mixers, and the advantages and disadvantages of such a circuit are well known. Therefore, only minimal effort will be spent in this area: that effort directed toward the determination of what stability can be expected from a backward-wave oscillator by the use of highly regulated power supplies.

2. Backward Wave Converter (BWC)

The ITT Laboratory, Nutley, N.J., is engaged in a program for

Patrick AFB (AF 08(606)-4660) involving the refinement and packaging of BWC and BWA tubes which will operate at frequencies from 500-12000 Mc. These tubes were originally developed for RADC (AF 30(602)2025). The major emphasis of the current program is wide dynamic range. The aim is for a dynamic range from -80 dbm to +36 dbm.

a. General Description

The Backward-Wave Converter is a Traveling-Wave Tube which combines in one envelope a backward-wave amplifier, a backward-wave oscillator and a mixing structure. ITT uses a cascade BWA section in order to improve stability and increase the reverse attenuation. The BWC operates as follows: the electron beam is common to the two sections of the BWA and the BWO, and actually passes through the BWA first. RF input signals modulate the beam in the BWA section. The beam then passes into the BWO section. The BWO oscillates at a frequency which differs from the center frequency of the BWA by the amount of the IF. Mixing takes place between the amplified RF input signal and the BWO signal in the beam. By placing a tuned circuit in the collector circuit, the IF output is obtained.

The BWA tubes are similar to the BWA section of the BWC.

b. Specific Points of Interest

(1) Choice of IF Frequency as Related to Image Rejection

The BWO section operates below the RF signal in all BWC units. Figures IX-10 through IX-14 show the selectivity characteristics of the X-389 BWA. These characteristics are typical of the BWA section of the BWC. As the BWO frequency is moved further from the center frequency of the BWA, image rejection improves. It would first appear that this process could continue until a level of image rejection of perhaps 70 db or so were achieved. (Note that the cold insertion loss of the BWA section is about 50 db and the gain is 20 db (see Figure IX-15). Therefore, it is not possible to have an image rejection greater than about 70 db no matter what combination of BWA and BWO frequencies is chosen). Figure IX-16 shows typical image rejection of an early BWC. In general, it is not possible to separate the BWA and BWO frequencies on a percentage basis because of the bandwidth imposed on the BWO. A second problem which is brought on by the use of higher intermediate frequencies is that of tracking the BWO and BWA sections. As their frequency separation increases, the BWO section and BWA section operate on portions of their nonlinear curves which become increasingly dissimilar. ITT feels an improvement of 10 to 20 db in image rejection can be achieved if the IF is increased 15%. One of the reasons for developing the BWA tubes for use in conjunction with the BWC tubes is to improve image rejection by virtue of the band-pass characteristic of the backward-wave oscillators.

(2) Spurious and Harmonic Responses

Not much information is available on the spurious and harmonic responses. ITT feels that the second harmonic response is probably better than 40 db down for the BWC tubes.
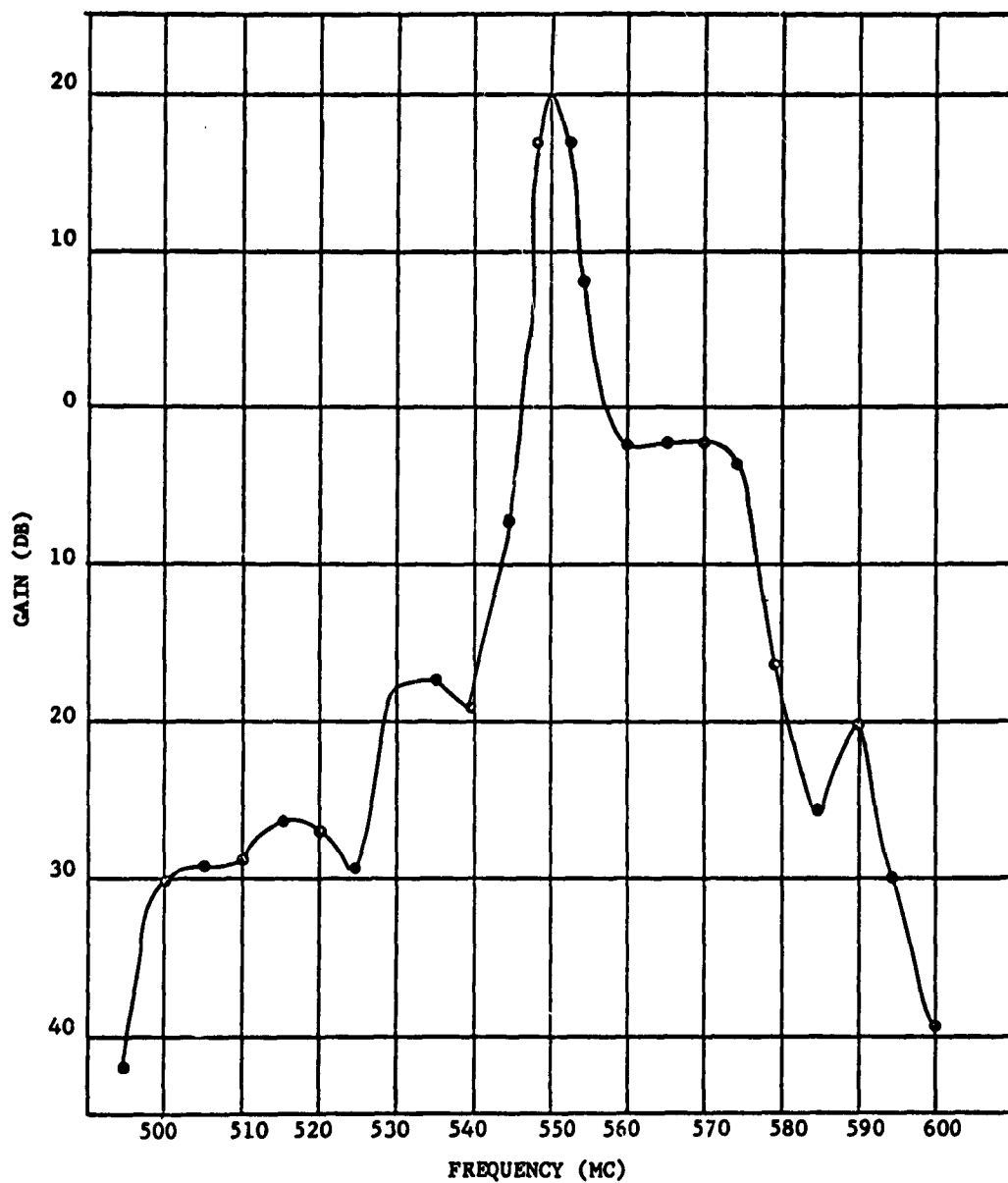
Figure IX-10.   X-389 BWA-Band Width When Tuned for 550 Mc
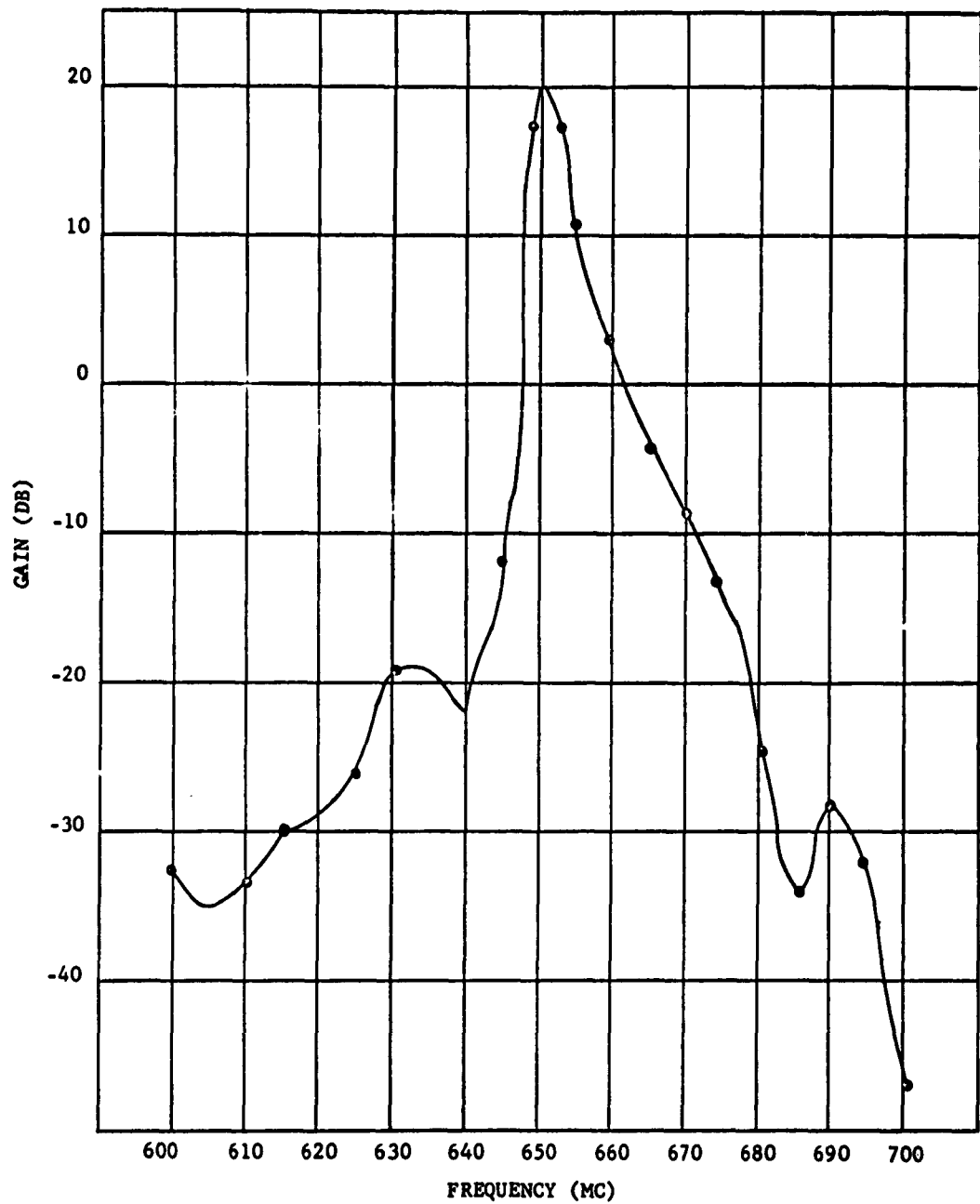
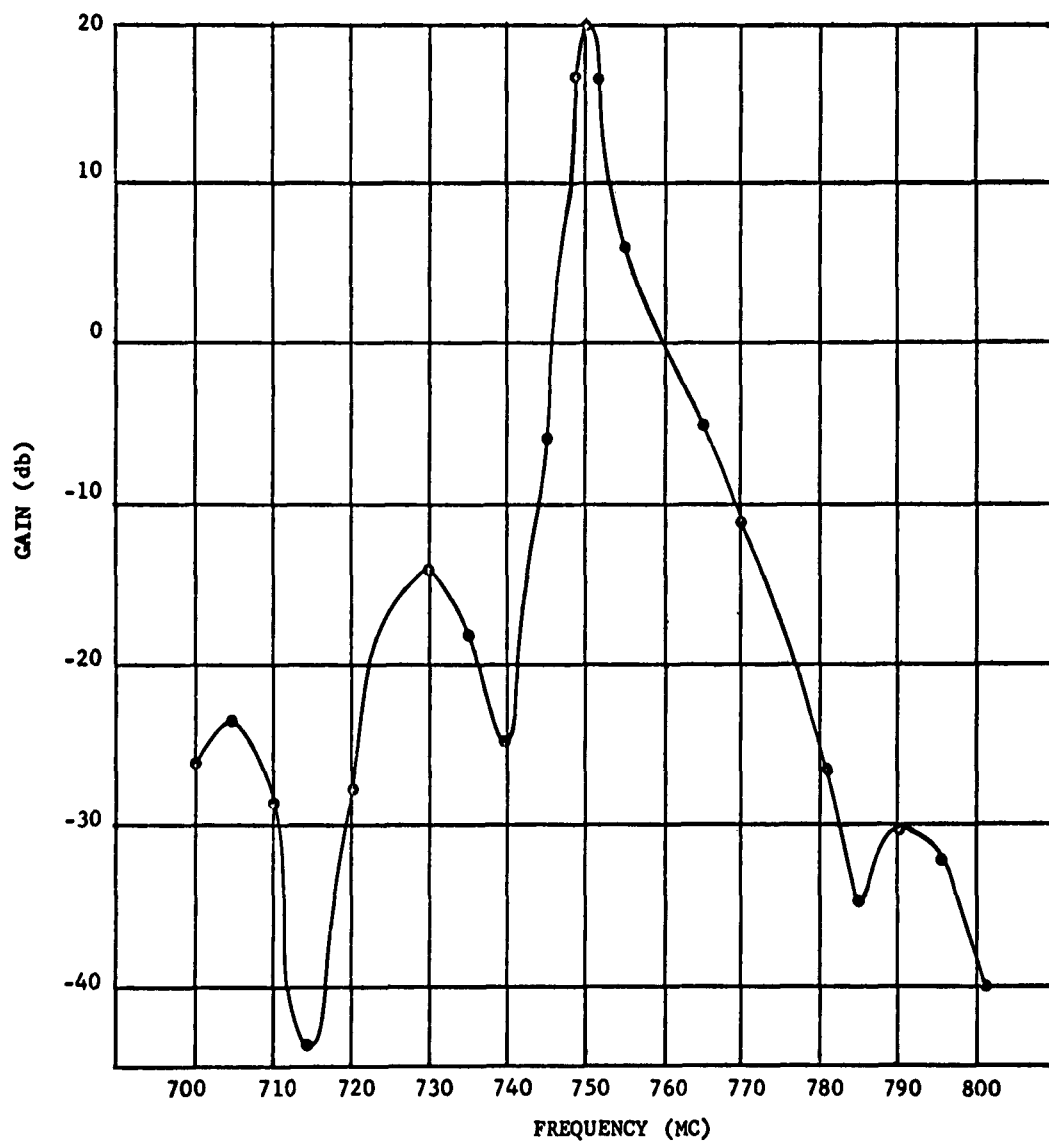Figure IX-11. X-389 Band Width When Tuned for 650 Mc

IX-18

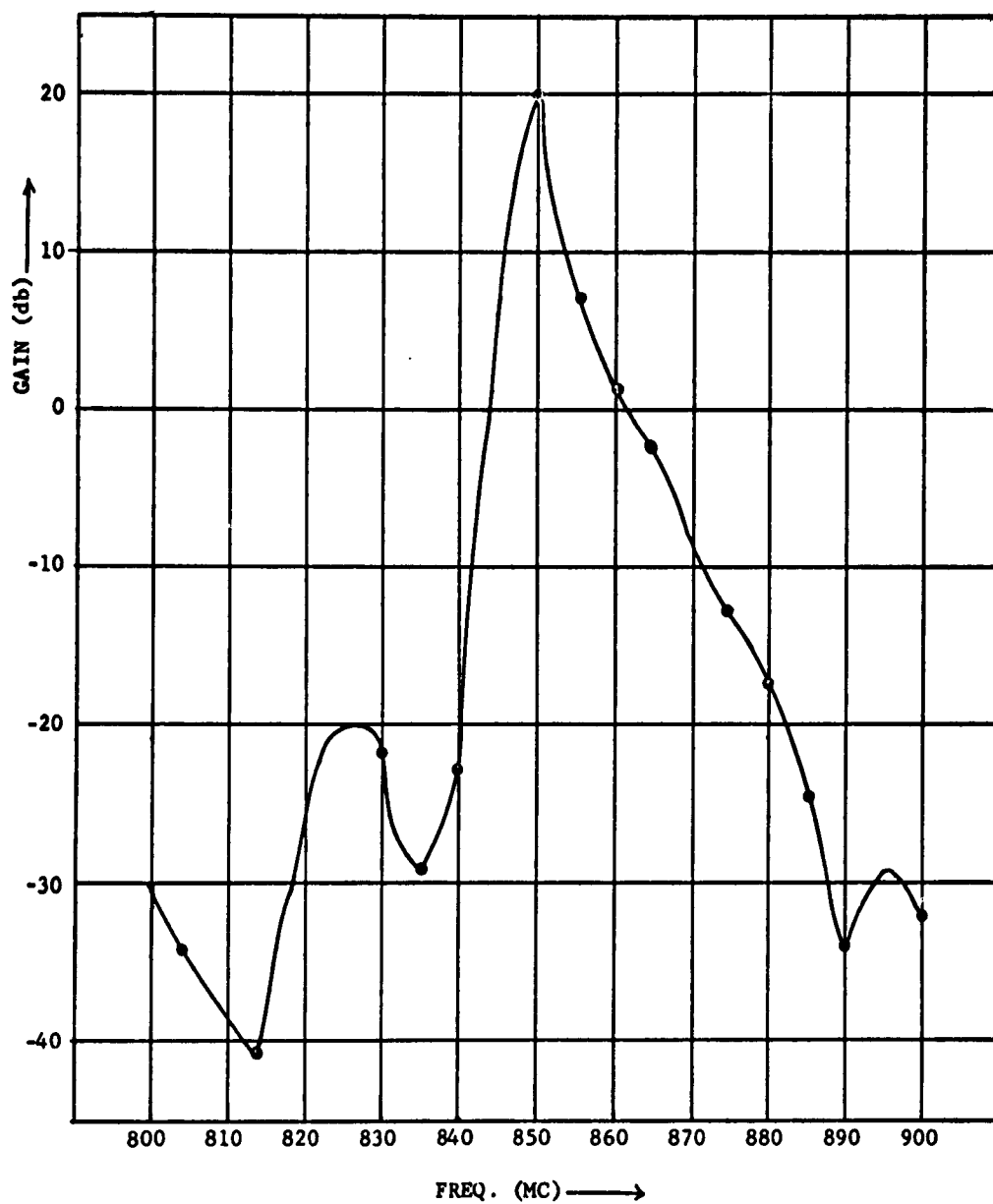Figure IX-12. X-389 BWA Band Width When Tuned for 750 Mc

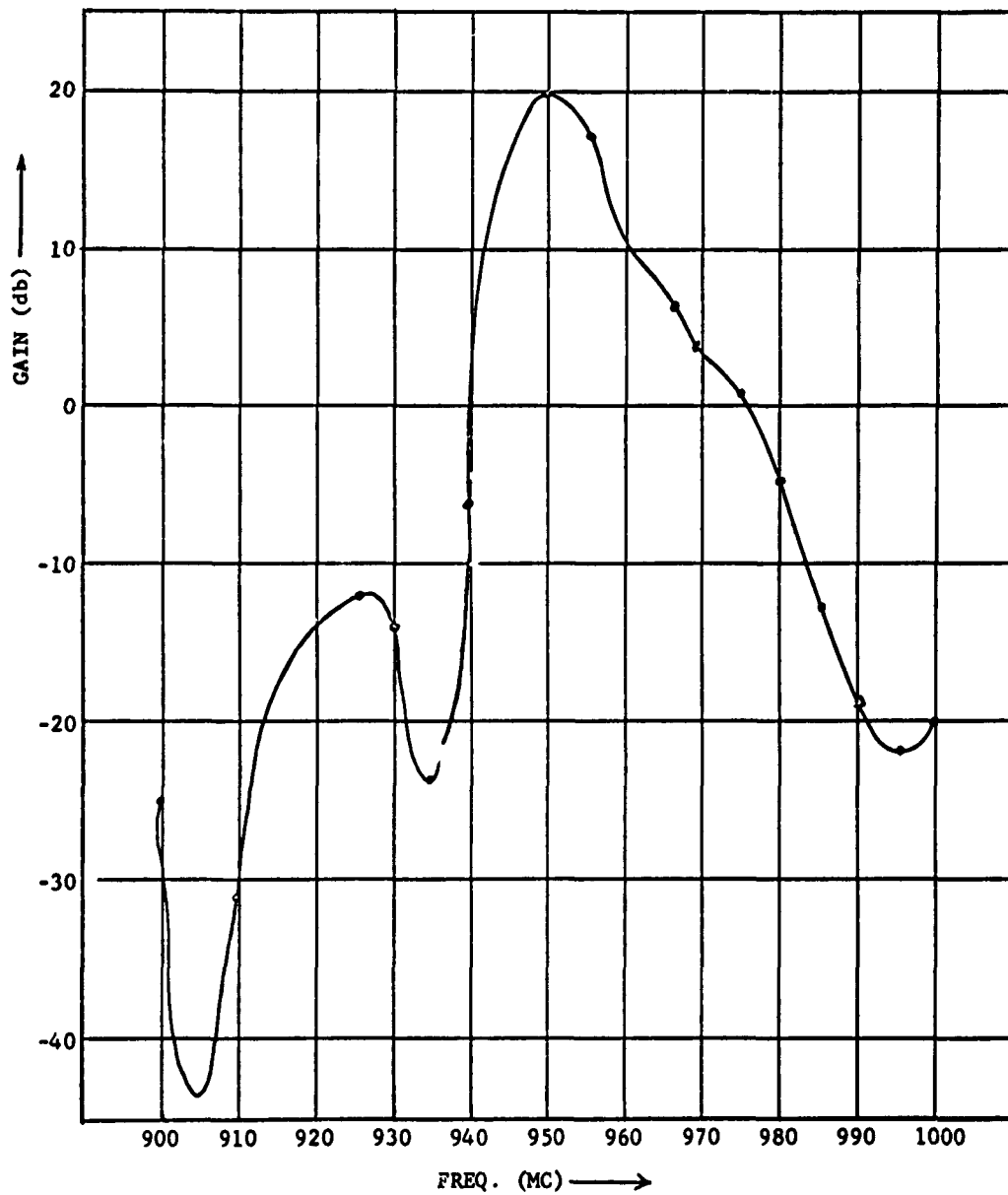Figure IX-13.   X-389 BWA-Band Width When to 850 Mc

IX-20

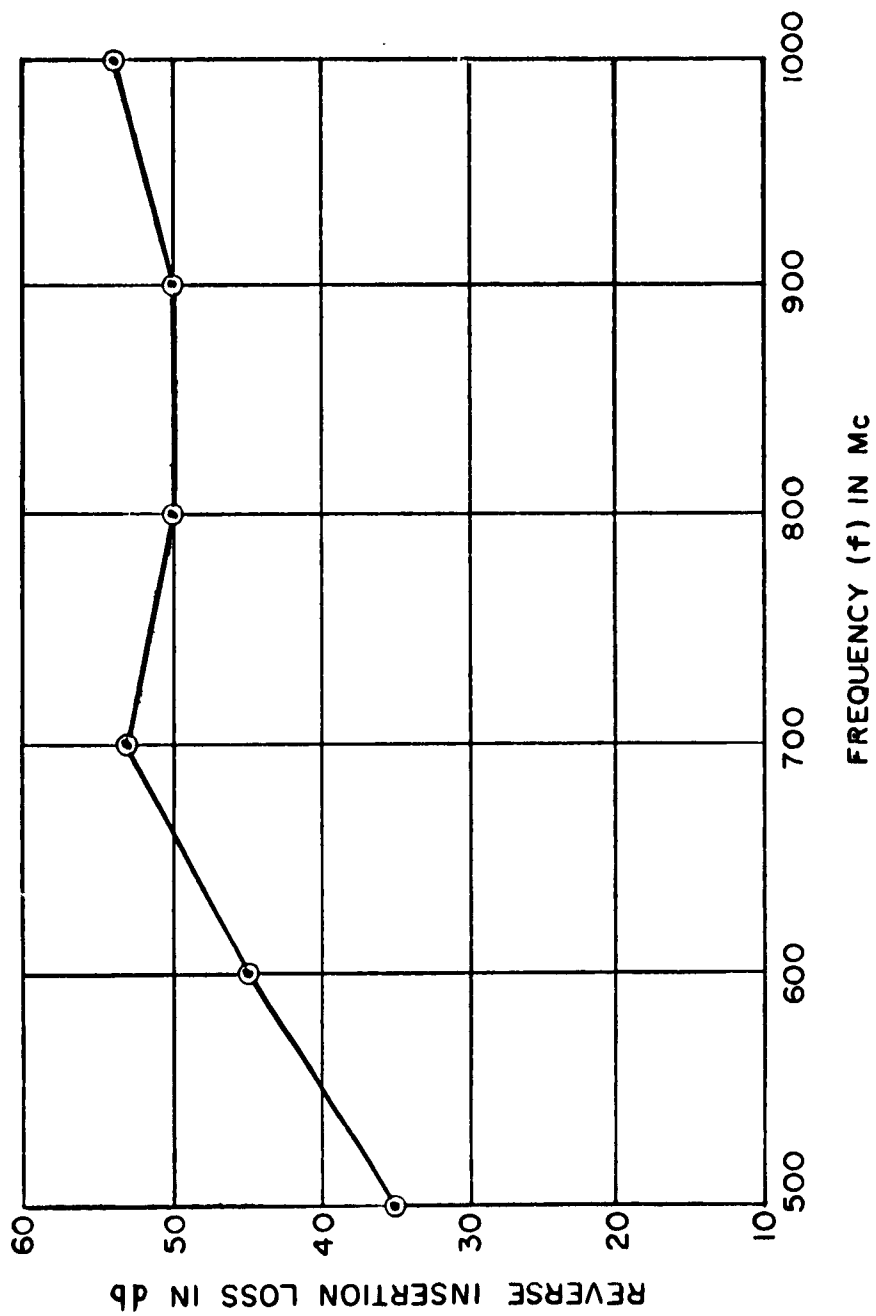Figure IX-14. X-389 BWA Band Width When Tuned for 950 Mc

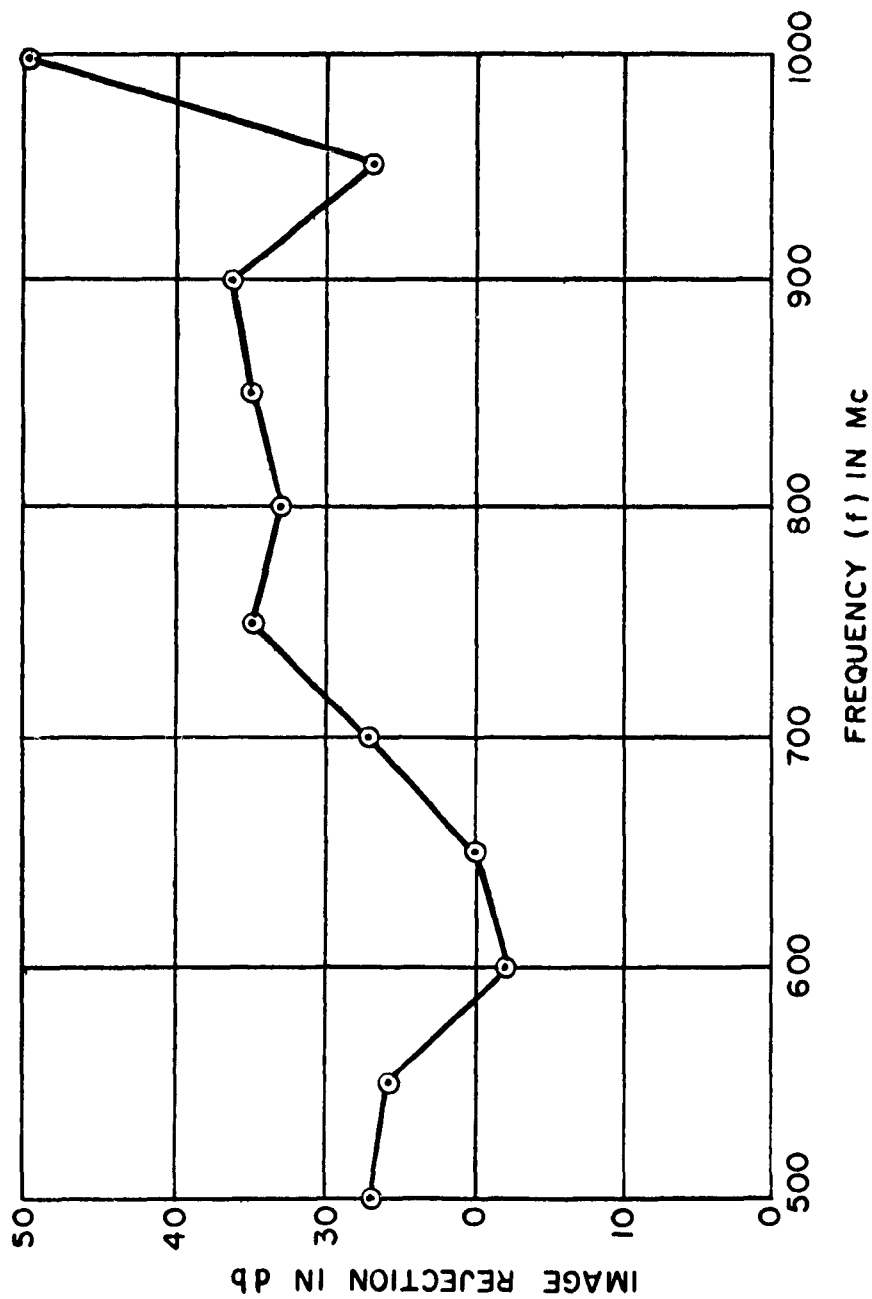Figure IX-15.   Reverse Insertion Loss Versus Frequency BWA

1101476

Figure IX-16.   Image Rejection Versus Frequency for the X-348 Backward-Wave Converter

1101477

(3) Sweep Rate Considerations

The lower the RF frequency, the longer the tube. This means that the maximum sweep rate is decreased. Rates of 100 kc in the X-band have been successfully used. However, in the 500- to 1000-Mc region, the maximum sweep rate is 1 kc. The biggest problem in sweeping is that virtually every electrode in the BWC must be swept. This includes, among others, all the electrodes making up the beam noise reduction profile.

(4) Noise Figure

Current noise figures are 15 db for the BWA tubes and 20 db for the BWC tubes. Figures IX-17 and IX-18 show noise figures in db versus frequency for an early BWA and an early BWC. Figure IX-19 shows the minimum discernible signal versus frequency for an early BWC. The noise figues of the tubes currently being built are high because the design goal for dynamic range is -80 dbm to +36 dbm. This makes large beam currents necessary. ITT feels that if beam currents were reduced, the noise figures for the BWC tubes would fall into the 10 to 15 db range (possibly closer to 10 db) and the BWA tubes would be about 10 db. One problem that this introduces is the requirement for longer tubes if the gain is to remain the same with reduced beam current.

(5) Tube Life

No life tests have been run. However, ITT feels that 1000 to 2000 hours can be reasonably expected. In general, the low frequency tubes will have a longer life than the high frequency tubes because of lower cathode loading.

(6) Environmental Aspects

These tubes are being built for ground based equipment. Because of this, tubes constructed of glass are baked at 400°C. Metal-ceramic tubes (which would allow higher baking and operating temperatures) are a possibility but might prove difficult because of input and output connection problems. The solenoids now used are water cooled. Air cooled solenoids are possible. PM focusing can be used on all tubes except the low-frequency tubes. PPM focusing structures are not feasible because of their incompatibility with voltage tuning methods and hollow beam configurations.

(7) Gain Flatness

The BWA's are programmed for a nominal gain of 20 db ± 3 db across the band. The BWC's normally have a conversion gain of -7 to +20 db depending on the Q of the output tuned circuit. For any given Q the conversion gain will vary less than ± 3 db across the band.

(8) Stability

The primary source of PM and gain instability in the BWC is the helix voltage. ITT feels that less than 0.01% ripple must be maintained.
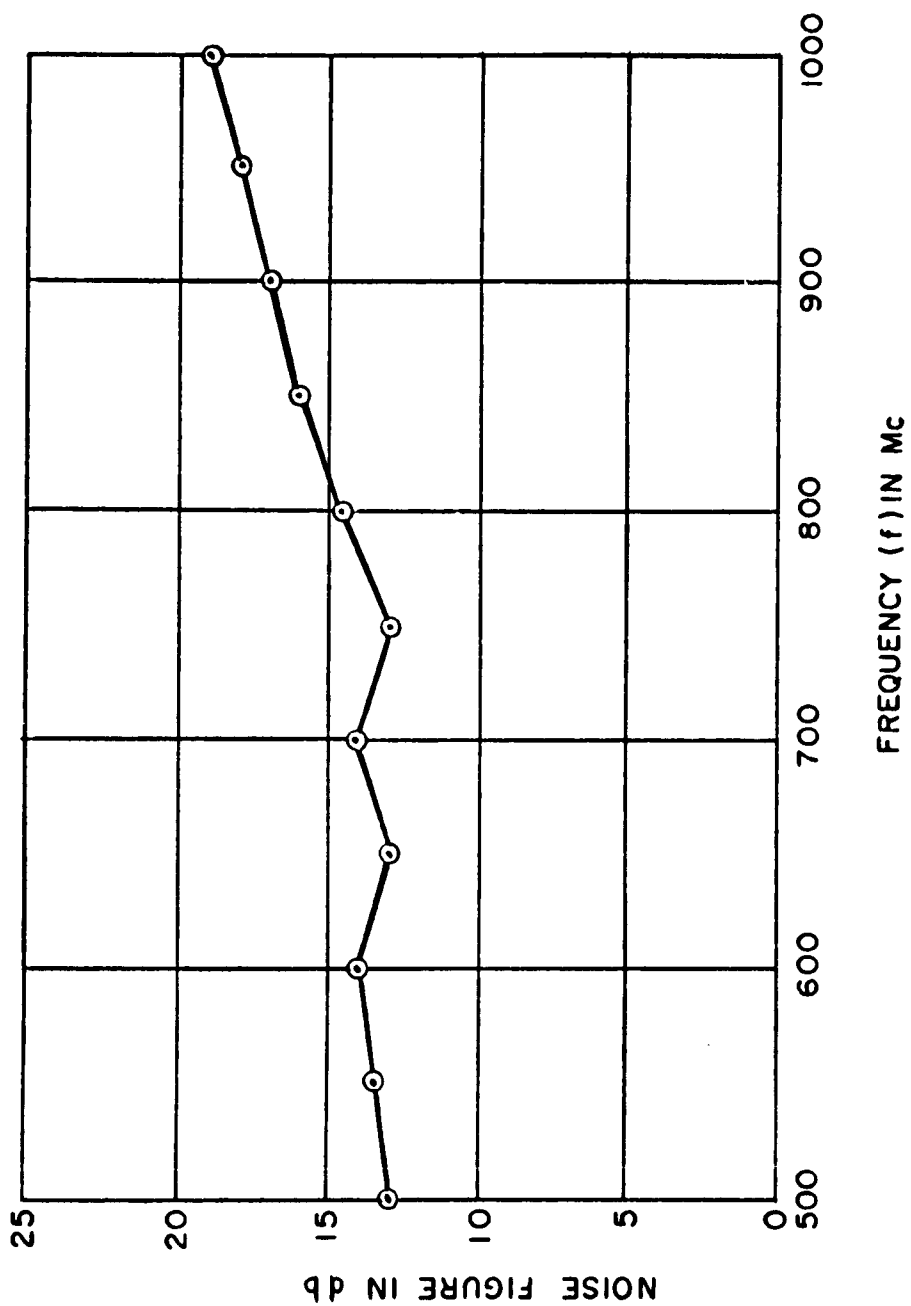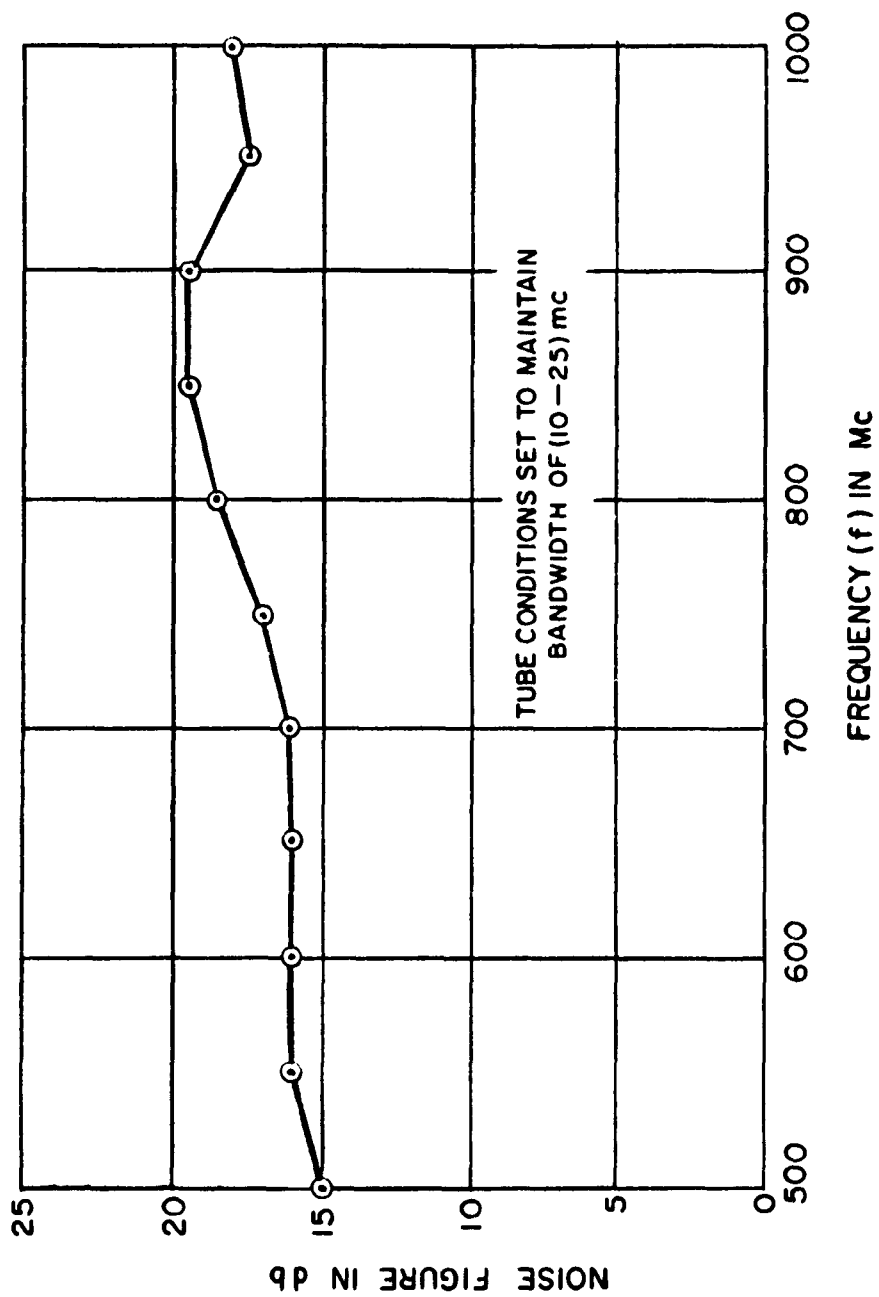
Figure IX-17. Noise Figure Versus Frequency BWA

1101478

IX-25

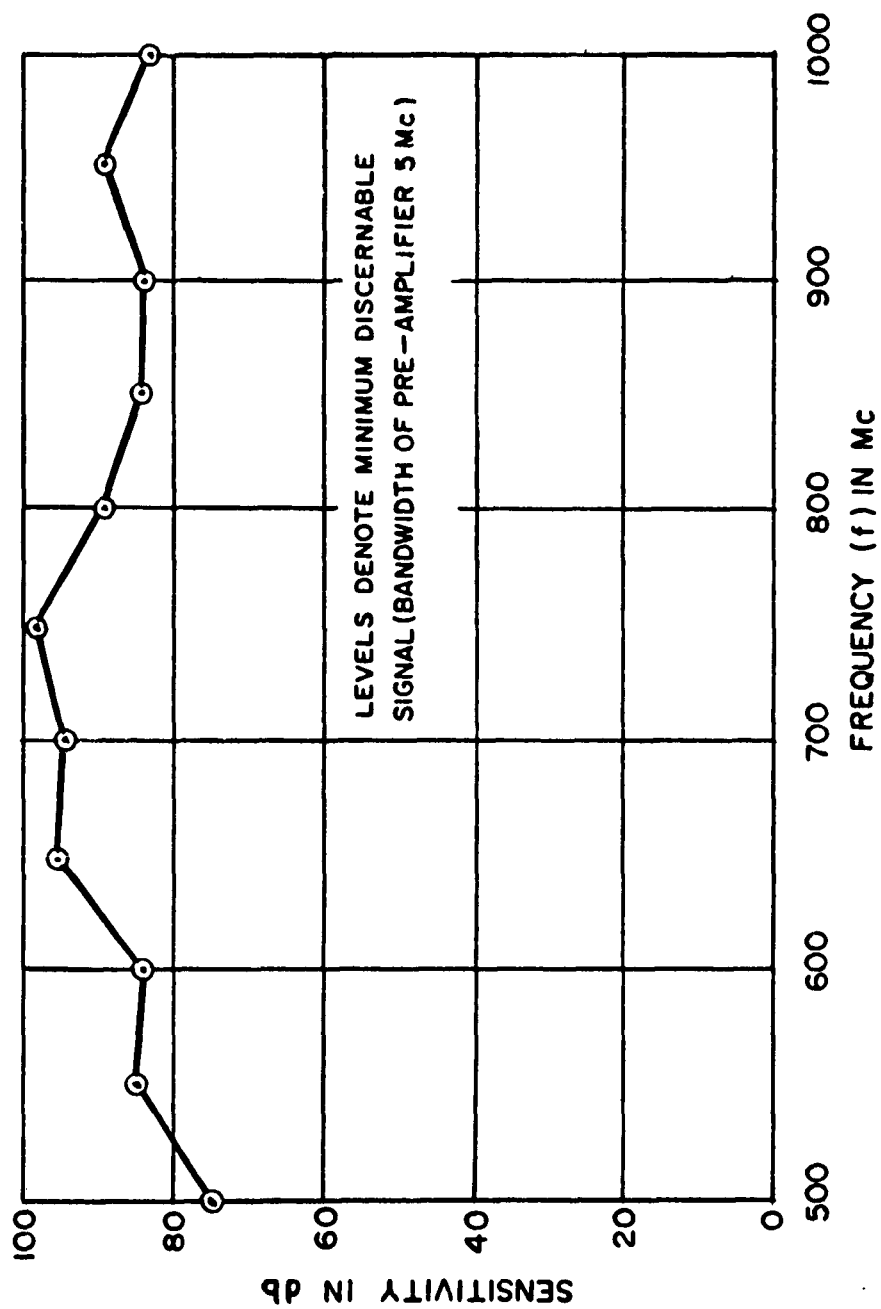Figure IX-18. Noise Figure Versus Frequency BW

1101479

Figure IX-19. Sensitivity of the X-348 Backward-Wave Converter Versus Frequency

1101480

The first anode is of secondary importance. Current regulation of 0.1% to 1% is required on the solenoids.

(9) Dynamic Range

As mentioned above, a dynamic range of -80 to +36 dbm is the design goal of the BWA-BWC system. In general, a better dynamic range is achieved at the high-frequency end of each tube's passband because of a higher saturated-power output. Figures IX-20 and IX-21 show the power output versus power input for the X-348 BWC. Figures IX-22 through IX-25 show the saturation characteristics of the BWC for the two-signal case.

c. General Comments

Several serious limitations are encountered when BWA tubes are used. The low-frequency tubes are out of the question for general receiver applications because they are too long. Since no method has yet been divised to bend the TWT beams around corners, a reduction in major dimension is not possible.

The noise figure (20 db) is excessive, the image rejection is too low (35 db) and the tubes are solenoid focused and require water cooling. Therefore, no further investigation of BWC devices is anticipated.

3. Tunnel-Diode Oscillator-Converter

a. General Discussions

The practicability of tunnel-diode local oscillators and down-converters over the 250-Mc to 40-Gc range is presently being investigated. The major characteristics desirable in a tunnel-diode oscillator-converter circuit are

> Low noise figure
>
> Large dynamic range
>
> Low intermodulation
>
> High frequency stability

Based on similar experience with experimental oscillators operating in the 1- to 2-Gc range, it is felt that the autodyne (down-converter) circuit offers the optimum compromise of the oscillator-converter criteria described above.

A hybrid-coupled mixer, employing two tunnel-diode mixers and a 3-db coupler, has been described in an article by Robertson.[IX-1] The main advantage of the hybrid circuit is that like the hybrid-coupled amplifier, it is stable without the use of nonreciprocal devices at the input port.

AEL is in the process of evaluating a self-excited down-converter. The only difference between the circuit of a tunnel-diode mixer,

---

(IX-1) W. J. Robertson, "A Broadband Hybrid Coupled Tunnel-Diode Down-Converter," Proc. IRE, V. 48, Nr. 12, pp. 2023-2024, Dec. 1960.
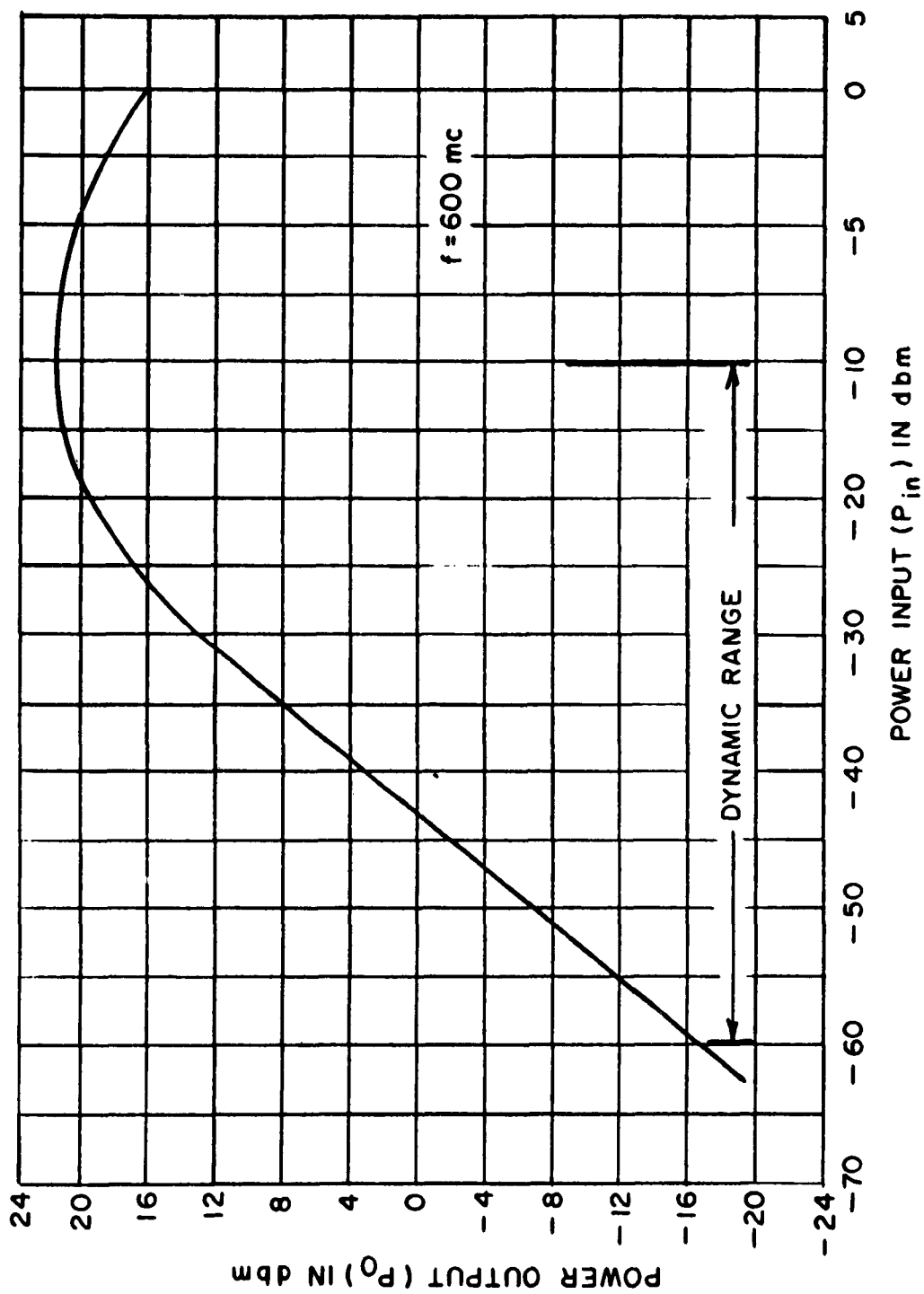
Figure IX-20. Power Output Versus Power Input for the X-348 Backward-Wave Converter
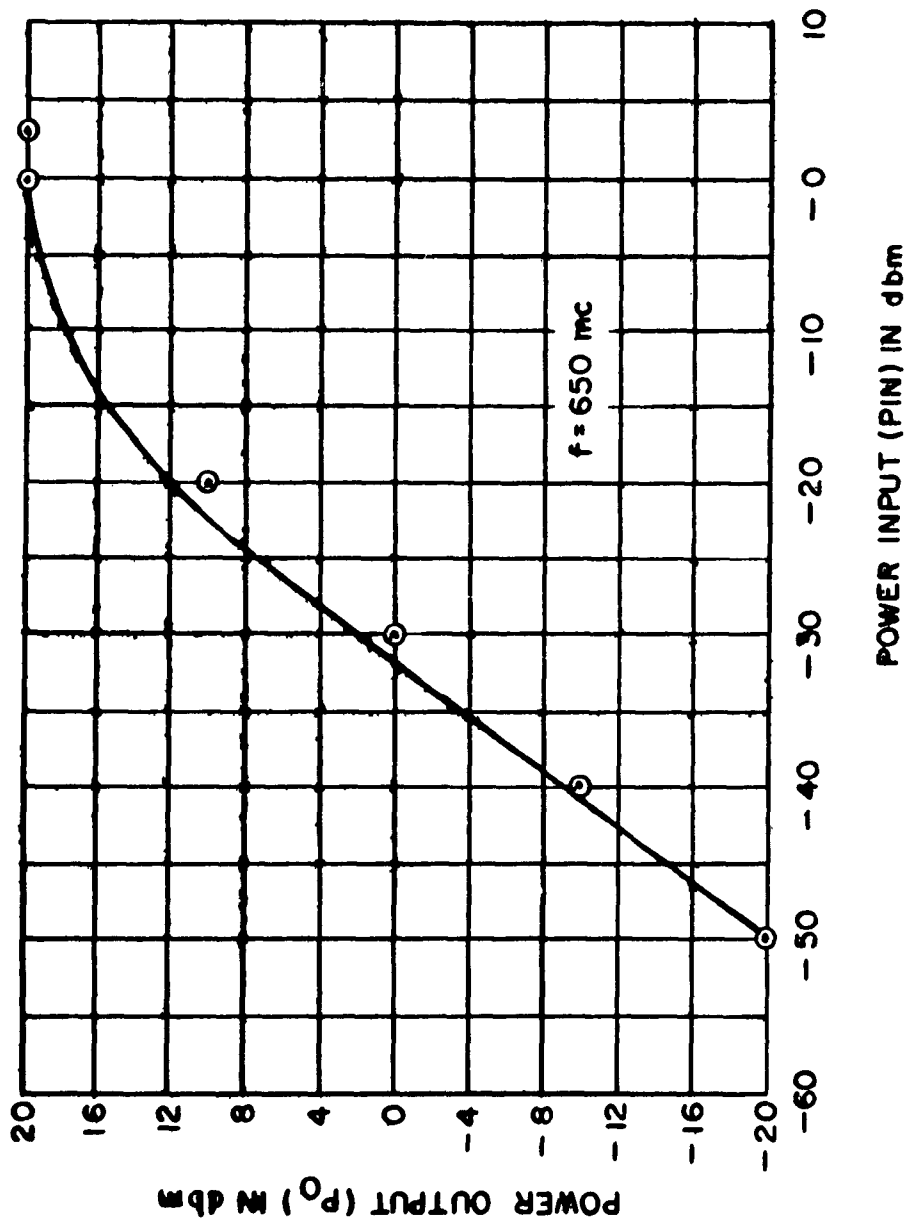
1101481

Figure IX-21. Power Output Versus Power Input
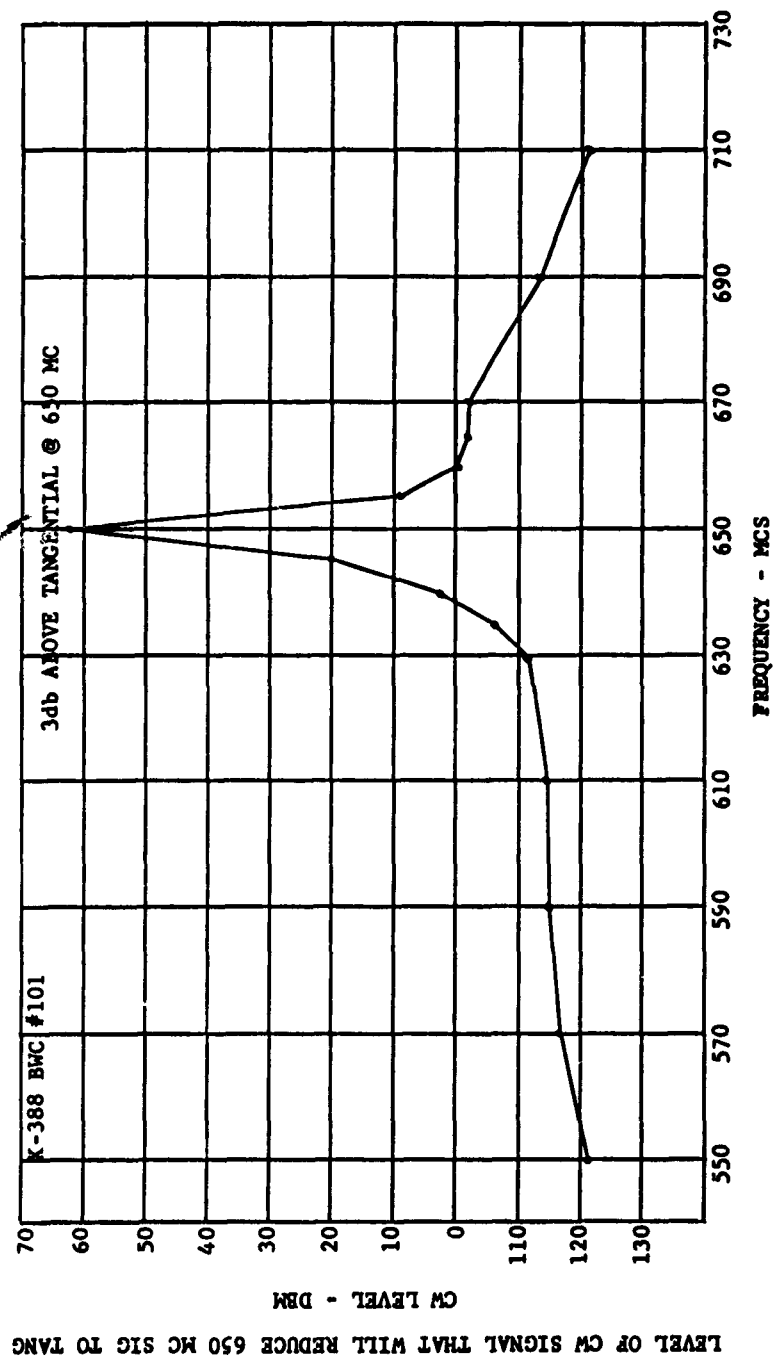for the X-348 Backward-Wave Converter

IX-30

Figure IX-22. Effect on Tangential Sensitivity by Surge CW Signal -
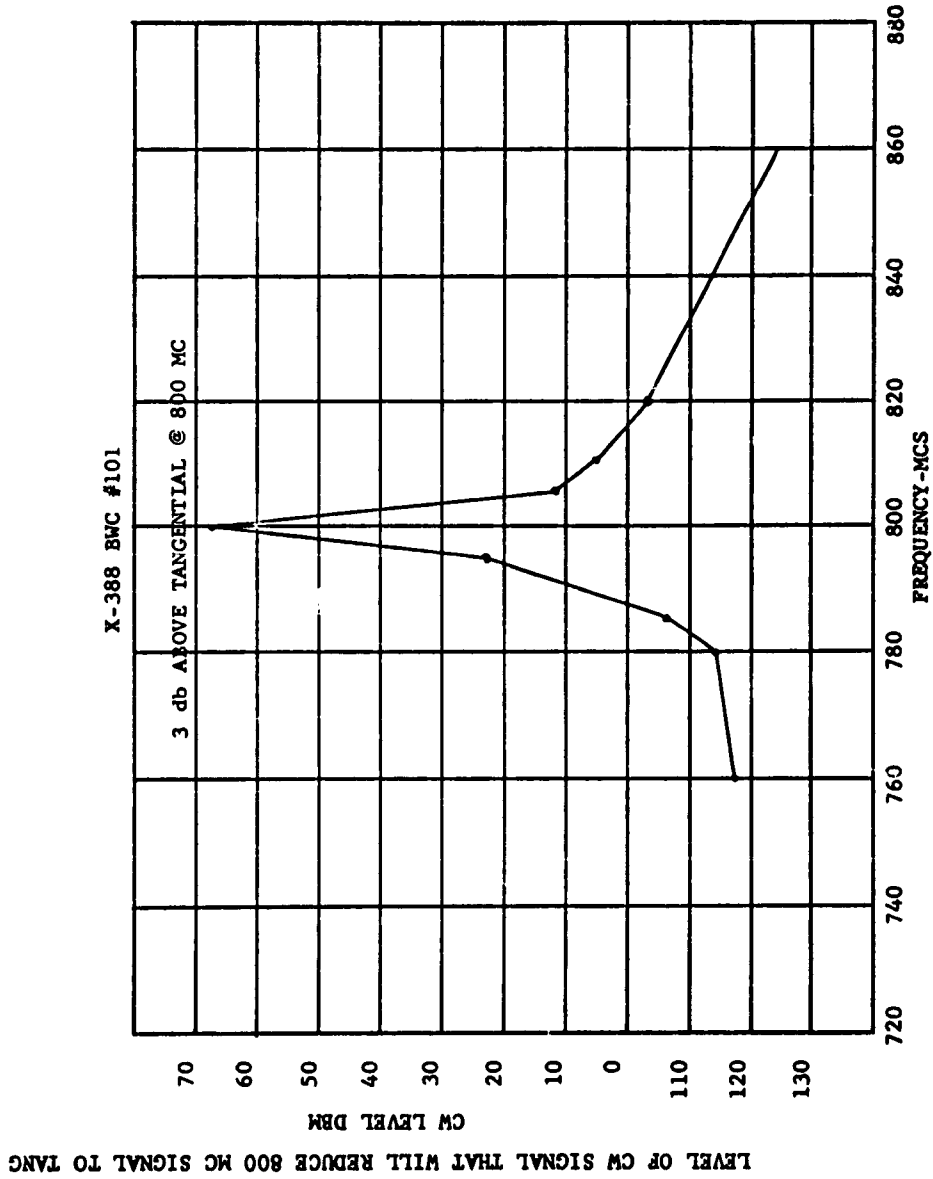X388-BWC 101

Figure IX-23. Effect of Tangential Sensitivity By
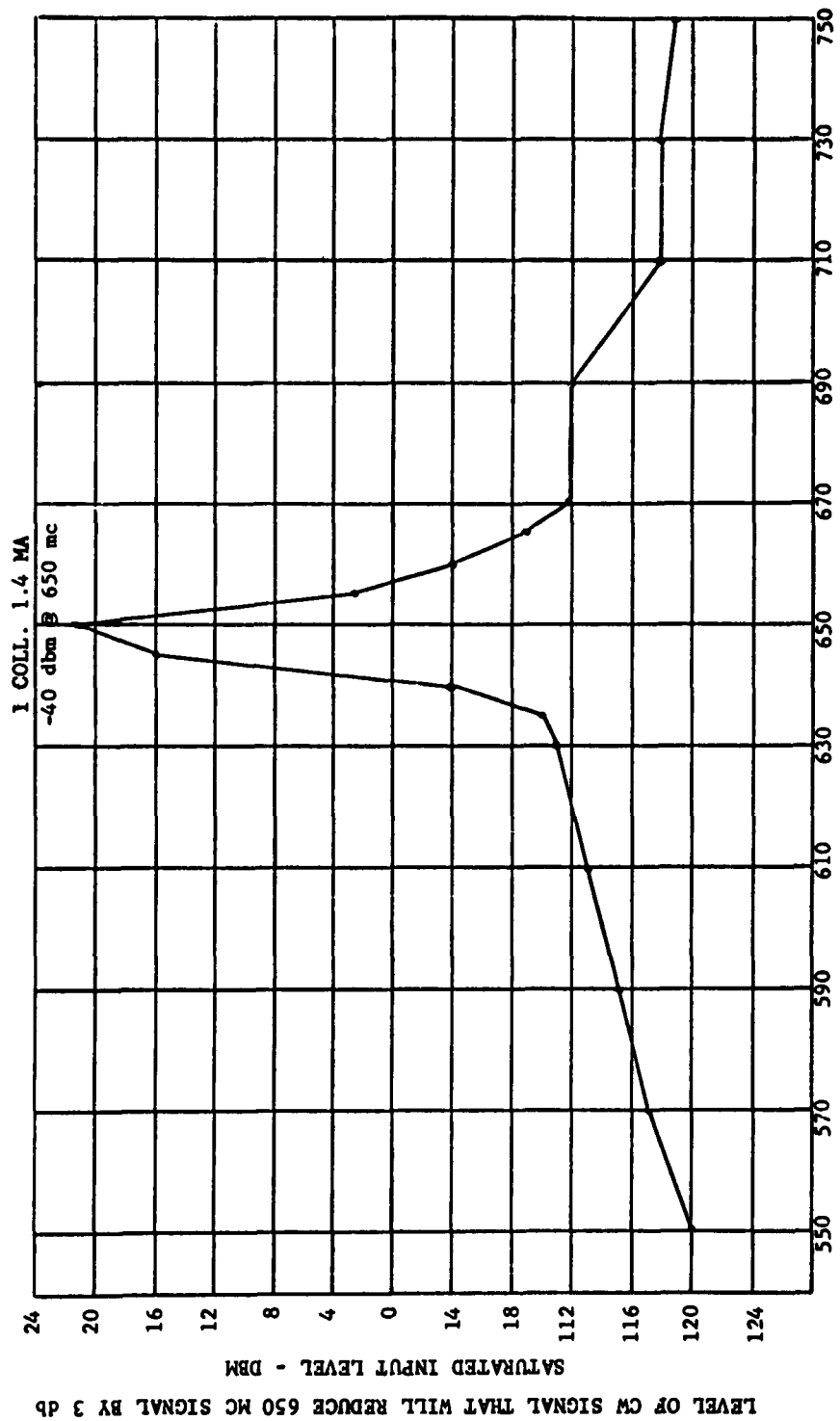Sarge CW Signal - X-388 BWC #101

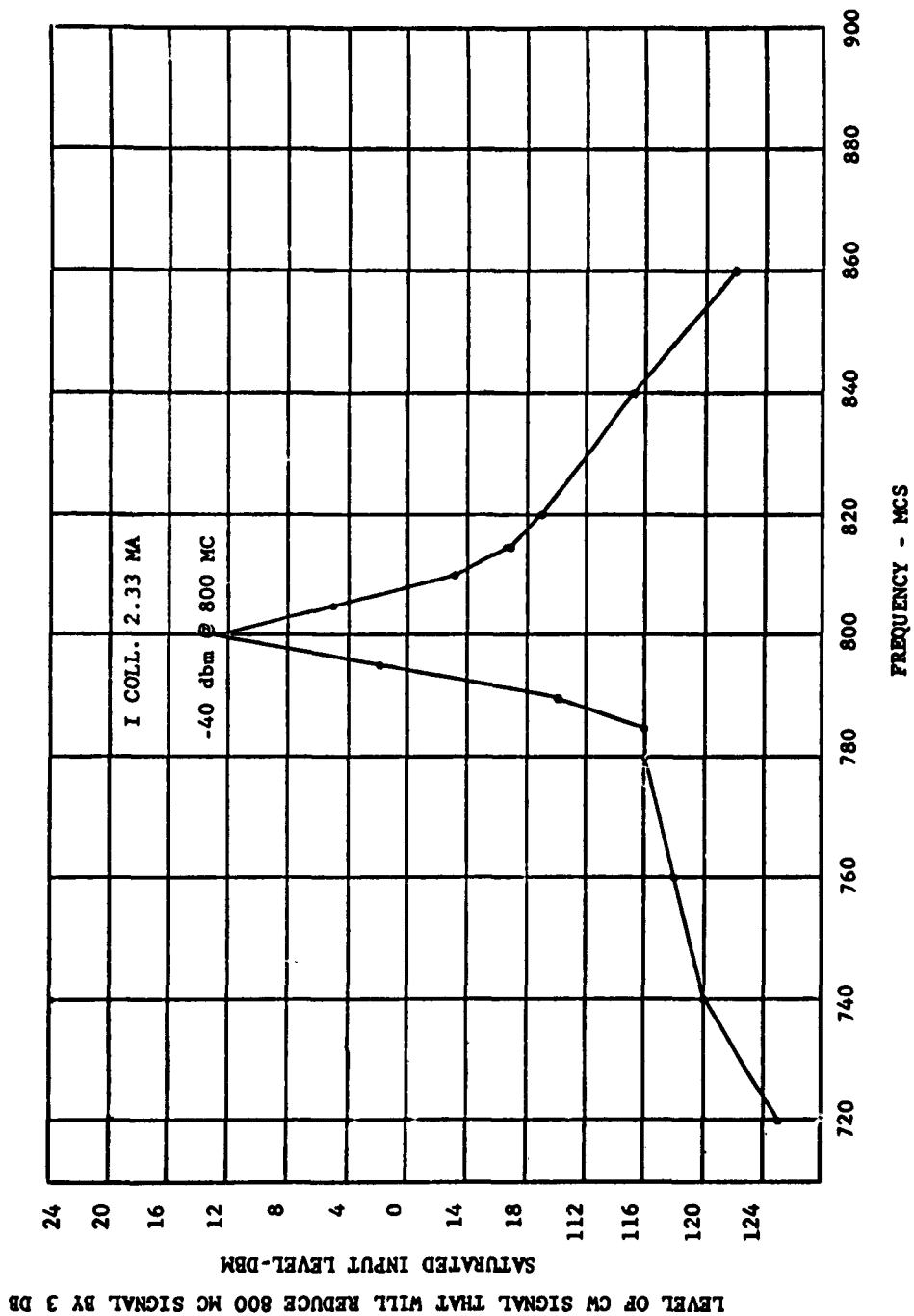Figure IX-24. Saturated Skirt Selectivity of BWA X-389 #103

Figure IX-25. Saturated Skirt Selectivity of BWA-389 #103

(Figure IX-26) and a self-excited down-converter is that the diode is biased far enough into the negative region for the diode to sustain oscillation.

The operation of down-converters up to 40 Gc will be investigated. Tunnel-diode oscillators operating at microwave frequencies up to 10 Gc have been reported in current literature. Fundamental power up to 40 Gc in frequency has been reported by Tramborulo and Burrus of Bell Labs.[IX-2] An output power at the low microwatt level and a tunability of a few percent is also recorded.

Based on our experience at frequencies up to 120 Gc, we feel that large, tunable bandwidths are feasible in the Ka Band. Oscillators utilizing specially constructed tunnel diodes can oscillate at 40 Gc in cylindrical cavities and by properly controlling case reactances, parasitic oscillation can be controlled and fundamental tuning can be achieved.

Various tunnel diodes have been used as oscillators with similar results. A list of these diodes include:

1. GE 1N3939

2. RCA 1N3138

3. Sylvania D4115A

4. Microstate M233

5. Microstate M234

During the next quarter of the program, four basic tunnel-diode oscillator parameters will be studied. These include (1) the diode itself, (2) the timing mechanism, (3) the bias technique and (4) the coupling method.

b. **Stability**

Frequency jumping, observed as a discontinuity in the tuning curve, is often noted in oscillators and down-converters whose cavity impedance is higher than 60 ohms. It can be eliminated if the cavity impedance is made less than 30 ohms.

Assuming the external inductance is much larger than the diode lead inductance, the frequency of oscillation of the shunt circuit is given by:

$$\omega = \left[\left(\frac{1}{L_c C} \right) \left(\frac{1 - R_s/R_D}{1 + R_s/R_L} \right) \right]^{1/2}$$

where: $R_D = \frac{1}{gD}$

(IX-2) R. Trambarulo & C.A. Burrus, "Esaki-Diode Oscillators from 3 to 40 Gc," Proc. IRE, V. 48, pp. 1776-1777, Oct. 1960.
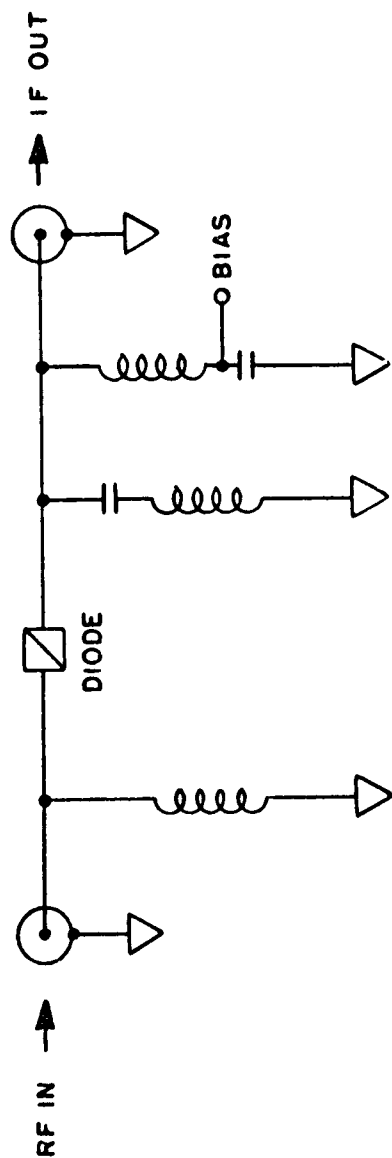
Figure IX-26. Typical Tunnel-Diode Mixer Oscillator

The requirement for stable oscillation is that $R_L > R_D$, assuming that $R_s \ll R_D$.

c. Sensitivity

By providing an IF output and changing the value of the bias voltage an oscillator can be made a down-converter. The tuning range of the down-converter is the same as the oscillator. The tangential sensitivity, of the down-converter, when limited to a 20 Mc IF bandwidth is about -90 dbm.

d. Noise Characteristics

The noise of a down-converter is not degraded by the ratio of the signal and the intermediate frequency as in the case of a parametric down-converter. The noise figure of tunnel-diode down-converters agree very closely to that of the same unit used as an amplifier.

One problem does exist when a down-converter is used instead of a separate local oscillator and mixer. The down-converter not only produces sinusoidal self-oscillations of the fundamental, but also produces harmonics of the oscillation frequency in the output. This problem can be reduced by operating the down-converters self-oscillations over the most linear region of the I-V curve.

The conversion gain and noise figure of the mixer are dependent upon the negative conductance and the conversion conductance. This requires the local-oscillator voltage to be large compared to the signal voltage. Therefore, a sacrifice in dynamic range is required in order to achieve an optimum noise figure in the down converter.

APPENDIX X

SECTOR SCAN FOR INDICATING DEVICES

In this appendix an analysis will be made to demonstrate the reason for the choice of the particular method of sector control to be demonstrated. Several schemes of sector scan control are presented. A brief mathematical analysis of improved Intercept Probability of sector scan control is also presented.

A. CONTROLLING THE SECTOR

When the operator desires to observe a narrow segment of the total scan in order to determine rapidly whether a signal actually is present, perhaps the most convenient control would be one which provides one knob to adjust the low frequency limit of the scan, and another to adjust the upper limit of frequency scan. In this way, he could closely observe the frequency at which a suspected signal exists while trimming away other portions of the frequency band. This type of control would present some problems when the provision of a calibrated expanded sector scan is required.

Another method which requires rather simple circuitry and provides a readily controlled sweep is proposed. The block diagram is presented in Figure 7 of the text. With this system the operator zeros in on a signal by first tuning the Center Frequency Control to the estimated frequency of the signal, then narrowing the sector scan to the desired amount. To facilitate the calibration of the indicator it may be best to change the sector by means of a switch rather than by means of a continuous control. This would facilitate calibration of the Indicator. If an additional trimming control were to be used, the graticle illumination could be extinguished when the trimmer is off its calibration point.

When the sector scanning is intended to resolve adjacent signals or to analyze the characteristic of a given signal, it is most expedient that the sector be expanded to fill the width of the indicator and be centered about a given center frequency with a control provided to widen or narrow the segment about that frequency, as desired by the operator.

The center frequency control will be calibrated to permit accurate determination of the frequency of a signal which has been centered in the sector. The need to narrow the sector to zero width requires the sweep circuits to be dc coupled throughout.

B. SEVERAL TYPICAL METHODS OF PRODUCING A SECTOR

Refer to Figures X-1 through X-6 which show typical waveforms and typical circuitry to provide different types of sector analysis in a panoramic receiver. The type shown in Figure X-3 is the type selected as being optimum for improvement of Intercept Probability and for ease of control operation. Figure X-1 shows a type of sectoring which utilizes clipping circuitry to confine the receiver tuning to the segment desired. The slicing diodes merely prevent the receiver from being
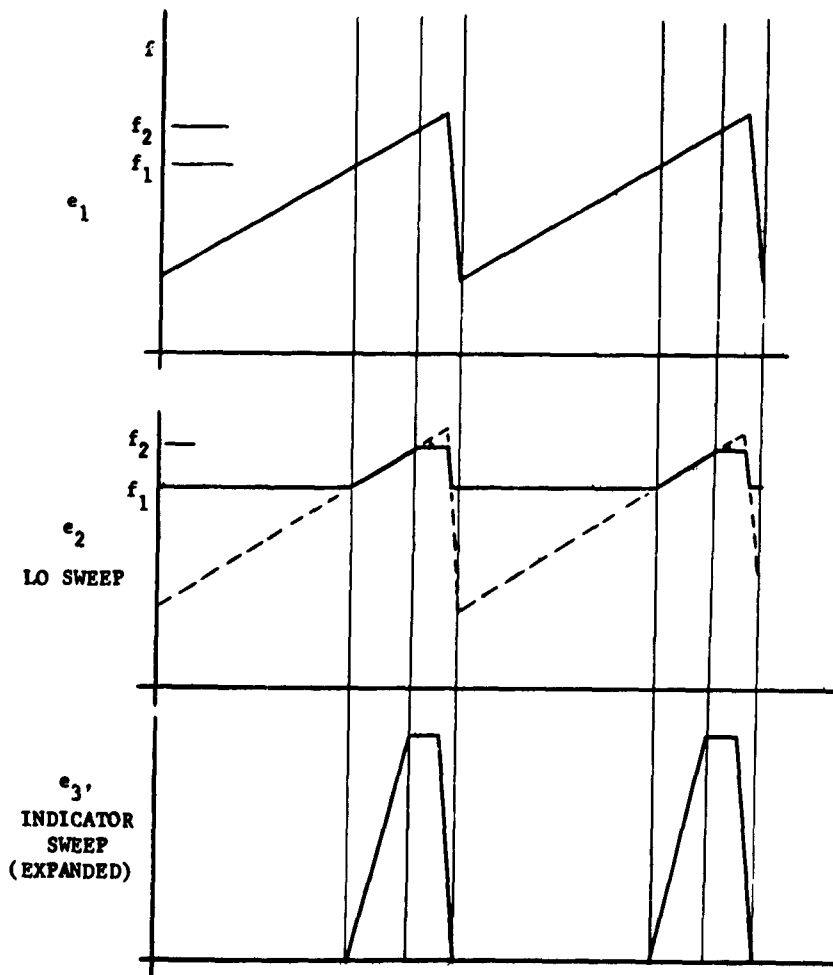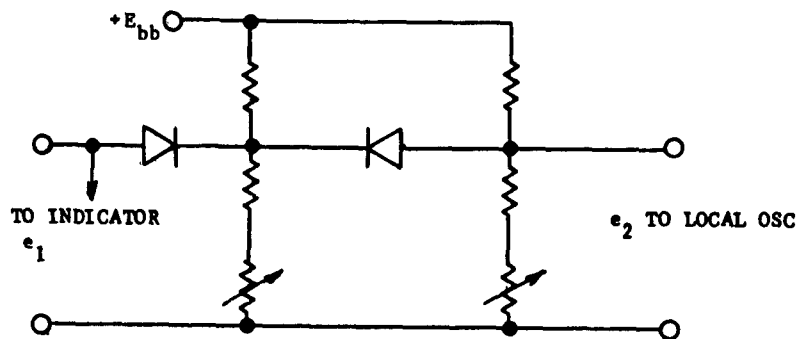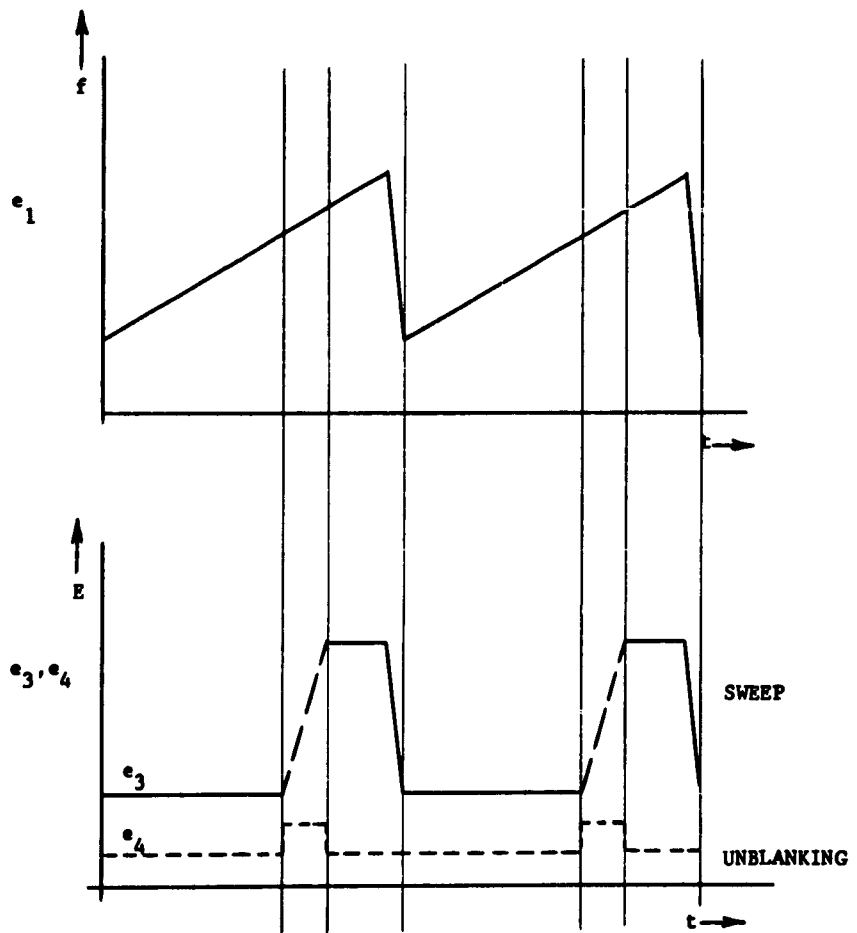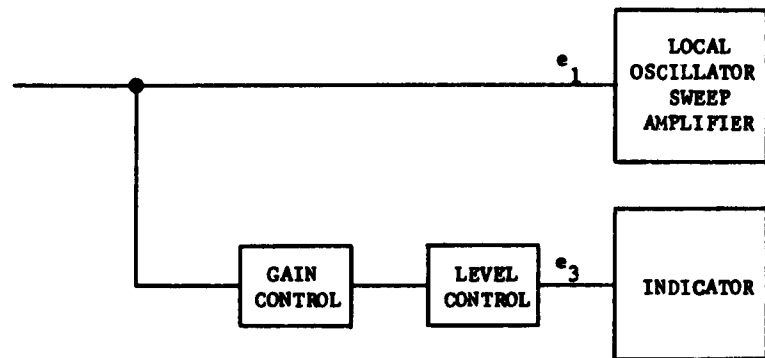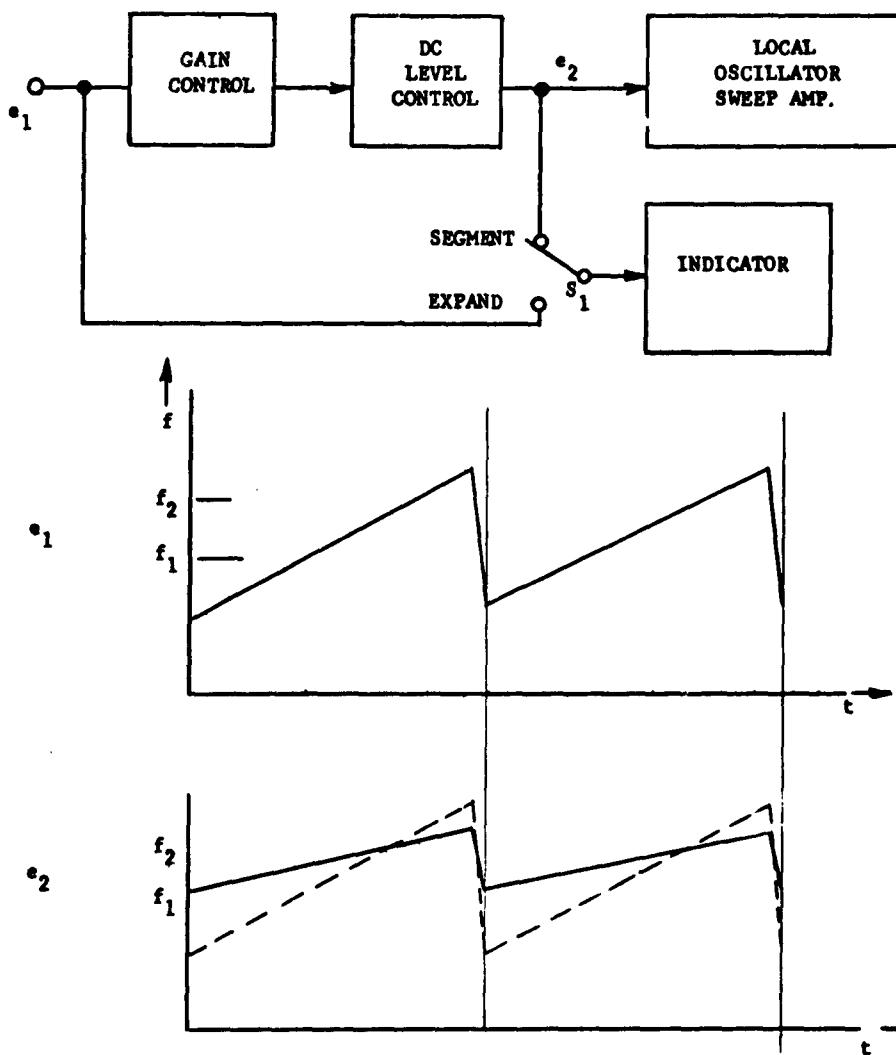
Figure FX-1. Sector Scan Method 1

Figure FX-2.   Sector Scan Method 2
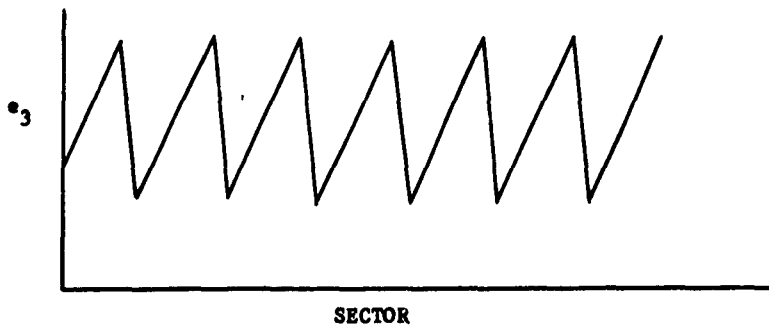
X-3

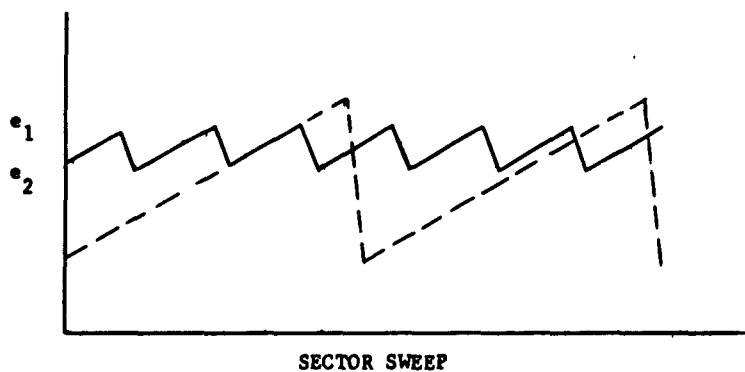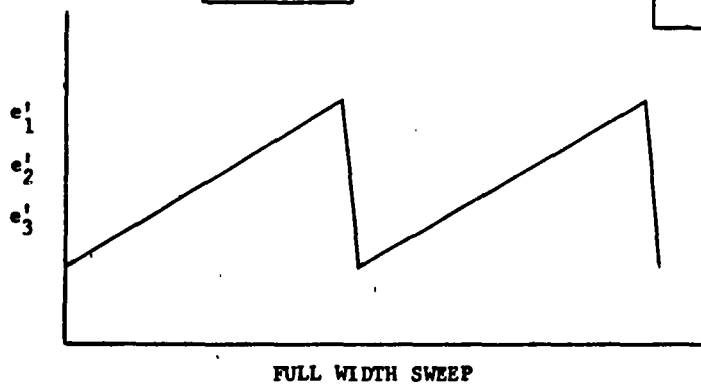Figure FX-3.   Sector Scan Method 3

X-4

FULL WIDTH SWEEP

SECTOR SWEEP

SECTOR

Figure FX-4. Sector Scan Method 4

X-5

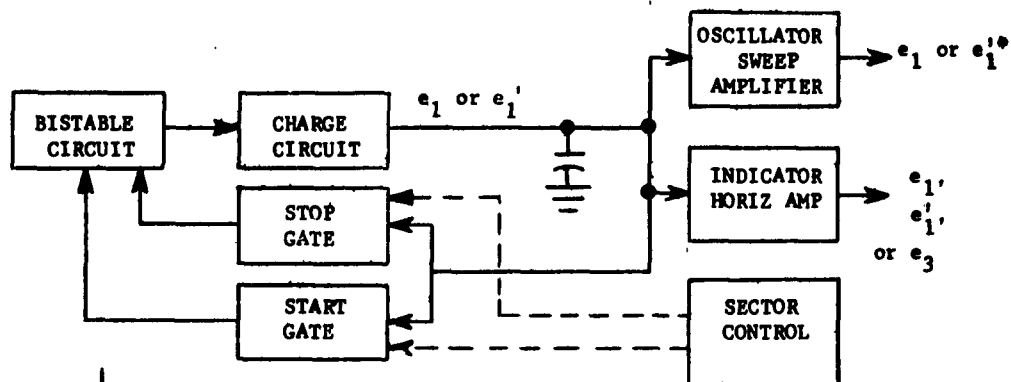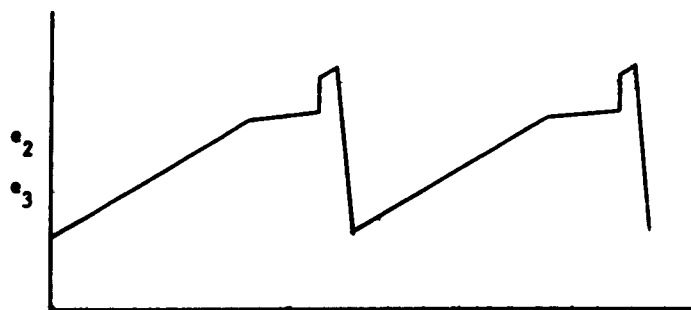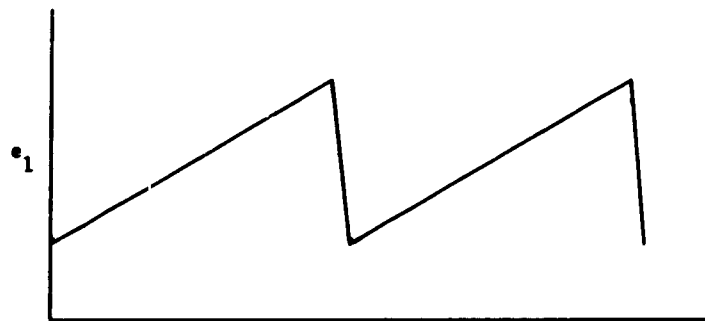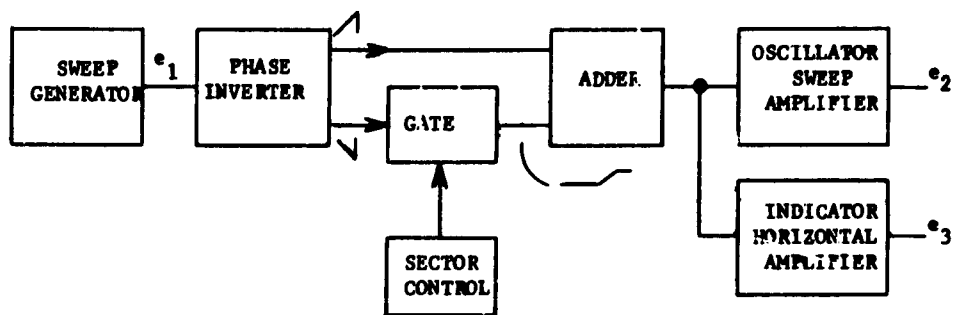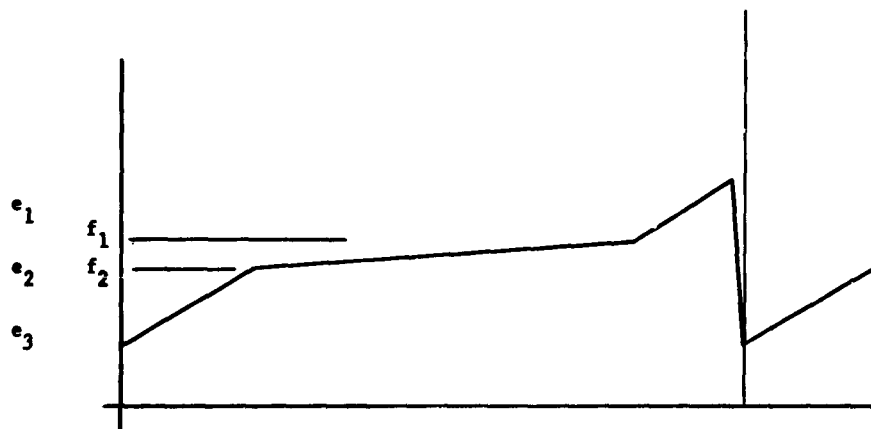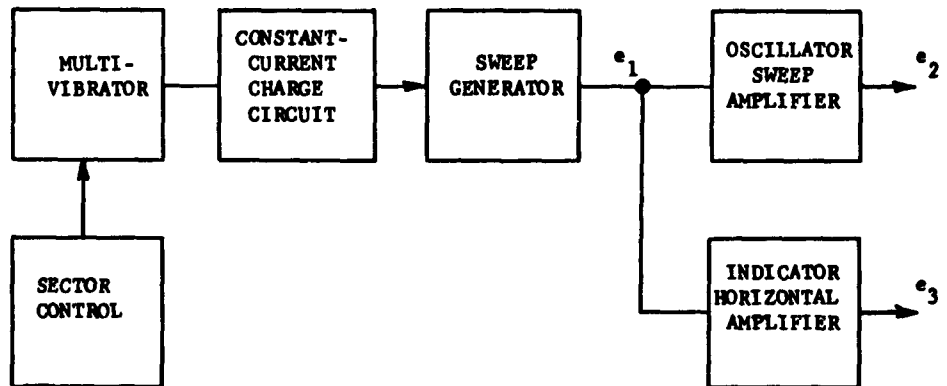Figure FX-5. Sector Scan Method 5

Figure FX-6. Sector Scan Method 6

scanned below a certain minimum frequency or above a certain maximum frequency. The voltage waveform e_1, shown, is the input voltage waveform that would be applied to the local oscillator and the indicator sweep for a full range of scan as is normal. When it is desired to sector scan, the waveform into the local oscillator would be set. If this same waveform, e_2, is applied to the indicator, the frequency segment will appear at its true location on the indicator scale. The acquisition interval for a given signal would be unchanged from the interval that would be obtained if the full scan of the receiver were being used. In addition, there would be problems in the calibration of sector scan. It may be seen that there is little advantage to scanning a segment in this manner and some more desirable types will be discussed.

Figure X-2 shows a method which may be considered to be an adaptation of the type used in Figure X-1. In Figure X-2 the local oscillator continues to sweep at the same rate regardless of the width of the sector being viewed all the way from full scan to the stop-sweep condition. Only the scan rate of the indicator and the DC position of the indicator scan voltage is varied. Again, as with the circuit of figure X-1, intercept probability is unchanged and, as before, calibrating the sector width would be difficult. The viewing of a sector in its true-frequency location would not be possible.

Figure X-3 shows the method of control which is the most expedient from many points of view. The sweep generator continues to generate a waveform of the same shape and duration regardless of the width of the sector being scanned. The amplitude and the DC level of that waveform is varied to control the range of the segment being scanned by the local oscillator. With a DC coupled indicator, the frequency segment can be displayed in its true location by merely sweeping the indicator from the same voltage that sweeps the local oscillators. The segment which is being scanned can be expanded to the full width of the indicator device simply by obtaining the indicator sweep from the waveform e_1 at the input of the sweep control amplifier. Calibration of the sector width gain control becomes very easy. This type of control is the same that was originally recommended in AEL proposal 62-1 which resulted in the award of this contract. The working out of a convenient set of controls which would allow the operator to trim down his sweep at each end would require some extra effort with this type of control and mechanical linkage would be required.

Both acquisition probability and resolution are easily improved with this type of control since narrowing the sector scan slows the sweep rate in megacycles per second and allows a longer look as the receiver passes over each signal frequency.

Figure X-4 shows the type of sweep, where, as the sector is narrowed, the sweep rate of the scan in megacycles per second remains a constant but more sweeps occur in a given time. Again the indicator presentation can display true frequency information if the indicator sweep parallels the local oscillator sweep. As with figure X-1 and

figure X-2, here also a problem exists in calibrating an expanded presentation of a segment. It is necessary to cause the indicator to scan at full width at varying sweep rates. This might be accomplished by gain controlling potentiometers and a mechanical linkage or by some form of automatic sweep amplitude control. Waveform control for driving an inductive tuning device would be more difficult with this type control than that in figure X-3.

Figures X-5 and X-6 introduce a different concept of sector scan. Here, the receiver is swept at different rates over different portions of its frequency coverage. The effect on the indicator is one in which the scan proceeds normally across the indicator then slows down over a portion of the sweep to provide better acquisition and resolution, then follows one of two actions. In the circuit of figure X-5, the receiver frequency will leap up to that frequency where it would have normally scanned, had the sector not been slowed. This does give a disadvantage of overlooking certain frequencies. In the concept of figure X-6, the receiver proceeds to scan at its original rate once the chosen sector has been scanned.

If either of these two types of scanning were to be provided, figure X-6 would be the circuit to be preferred since no frequencies within the band are overlooked. These slowed sector techniques have the advantage that other frequencies outside the sector being analyzed in detail are kept under observation, rather than being ignored as in the case of other sector types which scan only a portion of the total available frequency range. Any new information appearing on the band or old information disappearing from the band would be noted by the operator. Since the same waveform would always be used for scanning both the indicator and the local oscillator, true frequency calibration would exist. There would be little gain in trying to expand the sector in this type of scan unless no other type of sector scan were to be provided.

MATHEMATICAL ANALYSIS OF IMPROVEMENT OF UNIT COINCIDENCE PROBABILITY - (UCP) BY SECTOR SCAN

A. GENERAL DISCUSSION

The effect that sector scanning produces on the Unit Coincidence Probability (UCP) can be determined. Using the equation (VI-I-11) of this report we find,

$$p^{(1)} = k \, \frac{b_r + b_s}{B_r} \cdot \frac{T_r}{T_s} + D_s \qquad \text{(X-I-1)}$$

where

b_r = receiver acceptance bandwidth

b_s = bandwidth of all components of signal (above receiver noise)

B_r = Range of center-frequency covered by the receiver

D_s = Duty-cycle of signal

k = factor of active receiver sweep

T_r = period of receiver sweep

T_s = period of signal $(\frac{1}{p\,r\,f})$

If we assign primed values to the parameters which may be affected by a change of the scanned sector, we obtain a new UCP:

$$\left[p^{(1)}\right]' = k' \, \frac{\left(b_r\right)' + b_s}{\left(B_r\right)'} \, \frac{\left(T_r\right)'}{T_s} + D_s \qquad \text{(X-I-2)}$$

The ratio of improvement of probability, A, would be,

$$A = \frac{\left[p^{(1)}\right]'}{p^{(1)}} = \frac{k'}{k} \, \frac{\left(b_r\right)' + b_s}{b_r + b_s} \, \frac{B_r}{\left(B_r\right)'} \, \frac{\left(T_r\right)'}{T_r} \qquad \text{(X-I-3)}$$

For this investigation it will be assumed that a constant receiver acceptance bandwidth is maintained when the sector width is varied, then the improvement ratio A of equation (X-I-3) reduces to:

$$A = \frac{k'}{k} \, \frac{B_r}{\left(B_r\right)'} \, \frac{\left(T_r\right)'}{T_r} \qquad \text{(X-I-4)}$$

and a loss of UCP actually occurs.

B. THE CASE OF FIGURES X-1 AND X-2

Using equation X-I-4, it can be shown that no improvement is obtained when the sector is controlled in the way set forth in figure X-1 or X-2 because, scaling from the curves in figure X-1:

$$A = \frac{k'}{k} \frac{B_r}{\left(B_r\right)'} \approx \frac{4}{19} \times \frac{11}{3} = \frac{44}{57} \tag{X-I-5}$$

and a loss of UCP actually occurs.

C. THE CASE OF FIGURE X-3

Applying the same methods, referring to figure X-3, we obtain

$$A = \frac{k'}{k} \frac{B_r}{\left(B_r\right)'} = \frac{19/20}{19/20} \frac{10}{4} = 2.5 \tag{X-I-6}$$

and an improvement of 2.5 occurs.

D. THE CASE OF FIGURE X-4

In this case, two different results will be obtained depending upon the relation of the flyback time to the charge time.

If the ratio of flyback time to sweep time is constant, $k' = k$, and $B_r/\left(B_r\right)' = T_r/\left(T_r\right)'$, the improvement is:

$$A = \frac{B_r}{\left(B_r\right)'} \cdot \frac{\left(T_r\right)'}{T_r} = \frac{T_r}{\left(T_r\right)'} \times \frac{\left(T_r\right)'}{T_r} = 1 \tag{X-I-7}$$

It must be remembered that UCP applies to a single sweep of the receiver. In the present case, the number of sweeps in a given time varies as $T_r/(I_r)'$ giving that many more tries in the given time. The improvement factor might then be assumed to be:

$$A' = \frac{T_r}{\left(T_r\right)'} \tag{X-I-8}$$

If, on the other hand, the flyback time is constant:

$$k = \frac{T_r - T_f}{I_r} \quad \text{flyback and} \quad k' = \frac{\left(T_t\right)' - T_f}{\left(I_r\right)'}$$

also, $\left(B_r\right)' = B_r \dfrac{\left(T_r\right)' - T_f}{T_r}$ and the change in number of sweeps

per unit time $= T_r / \left(T_r\right)'$

so: $$A'' = A\, T_r / \left(T_r\right)' = \frac{k'}{k}\ \frac{B_r}{\left(B_r\right)'}\ \frac{\left(T_r\right)'}{T_r} \tag{X-I-9}$$

$$A'' = \left[\frac{\dfrac{T_r - T_f}{T_r}}{\dfrac{\left(T_r\right)' - T_f}{\left(T_r\right)'}} \times \frac{B_r}{B_r\ \dfrac{\left(T_r\right)' - T_f}{T_r - T_f}}\right] \times \frac{T_r}{\left(T_r\right)'}$$

$$= \left(\frac{T_r - T_f}{\left(T_r\right)' - T_f}\right)^2\ \frac{\left(T_r\right)'}{T_r} \tag{X-I-10}$$

E. THE CASE OF FIGURE X-5

This type of sector scan presents a special case since sector analysis of part of the scan leaves unchanged the UCP and Acquisition Interval for signals outside the sector. Within the sector there is an improved UCP and at the right end of the sector where the sweep voltage returns to the normal value UCP reaches nearly to zero.

Equation (X-I-4) may be rewritten:

$$A = \frac{k'}{k}\ B_r/T_r \Big/ \left(B_r\right)' / \left(T_r\right)' \tag{X-I-11}$$

for the case under investigation $\left(B_r\right)'/\left(T_r\right)' \ll B_r/T_r$ within the sector, resulting in a large value of improvement, A, within that range. The value of k remains unchanged outside the sector, and does not apply within the sector.

F. CONCLUSIONS

Sector scanning shows an improvement of signal acquisition rate in nearly every scheme of sector scan presented, though it actually did show a loss in one case. The type of sector control presented in figure X-3 has been chosen for use in receivers to be demonstrated under the present study programs. The method easily demonstrates improvement of acquisition rate and demonstrates the ease of control which can be achieved.

APPENDIX XI

AN/GRR-9 FIELD TEST

A. GENERAL

The AN/GRR-9 receiver was developed by AEL, Inc. for RADC under
contract AF-30-(602)-1894. It was intended to be used primarily as
rapid-surveillance receiving system which would indicate the presence
of signal activity within the 1- to 10-Gc frequency range. The present
contract, AF30-(602)-2677, was initiated as a research study program
to develop techniques which would not only extend frequency coverage
of the receiver, but which would also develop receiver methods for
spectrum signature and field intensity measurements. Shortly after
the initiation of the present contract, it was deemed desirable to
conduct a field test wherein an existing AN/GRR-9, along with other
instrumentation would be used to obtain practical information and data
on typical field problems. This would serve in the development of
operational and equipment performance concepts under the new program.

B. SPECIFIC TESTS

The field test was divided into four separate tests. The objective,
procedure, results and conclusions of each test are described as follows.

   1. Test No. 1 - Transmitter Spectral Signature Indication

      a. Test Objective - To determine the effectiveness of an AN/GRR-9
         type system in intercepting and observing transmitter harmonic
         and spurious outputs.

      b. Test Procedure* -

         (1) AN/GRR-9, Serial No. 2, located in mobile van at Building
             No. 3 at the Verona test site.

         (2) AN/FPS-65 was radiating and rotating. An S-band and C-band
             radar were also operational.

         (3) All four GRR-9 antennas, vertically polarized, were mounted
             on the van roof.

         (4) All four GRR-9 displays were optimized.

         (5) Crystal video detection (characterized by large signals
             "running" across the indicator screen) resulted from the
             overloading of the L-band receiver by the AN/FPS-65 signal.
             A 50-db attenuator was placed between the L-band antenna
             and the receiver to eliminate saturation.

      c. Test Results

         (1) Test No: 1

---

* Refer to Figure XI-1 for a block diagram of the AN/GRR-9 receiver for
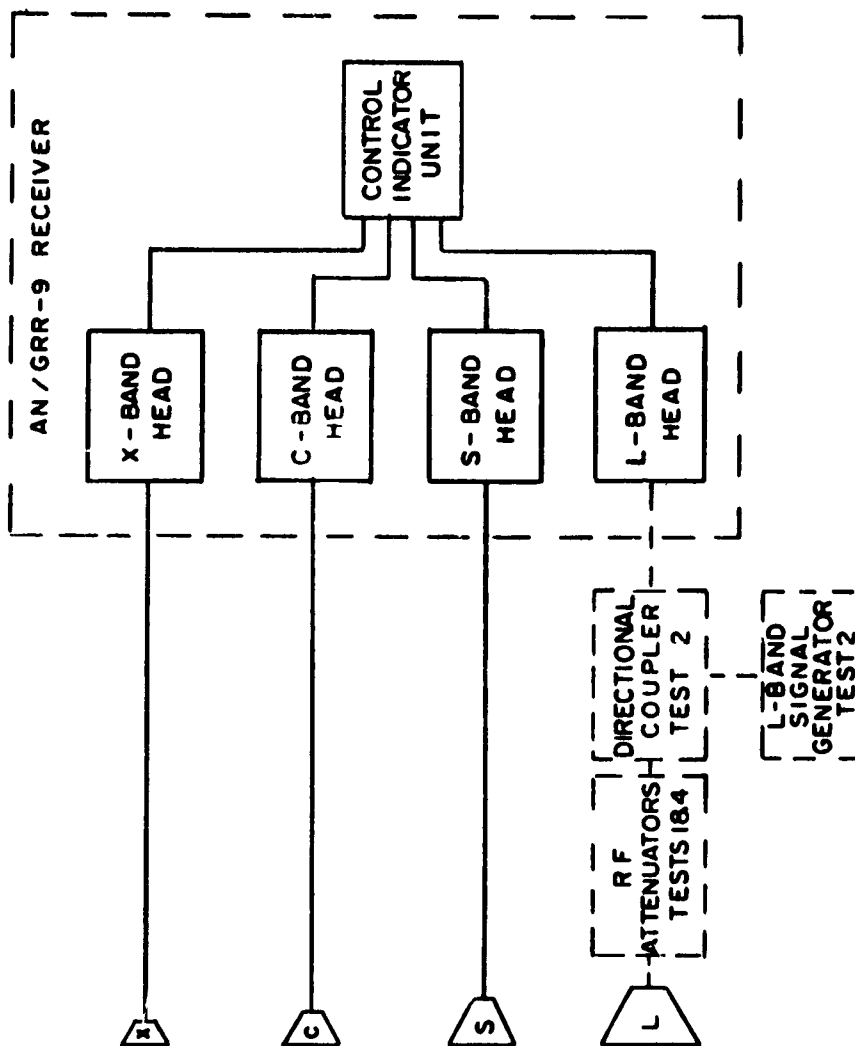 each test.

Figure XI-1. AN/GRR-9 Test Arrangement

1101431

XI-2

(2) Date: 10/17/62

(3) Recorded By: W. Brown, AEL

(4) Observers: A. Sparacino } RADC
    A. Jackson   }

The fundamental and harmonic outputs from the AN/FPS-65 were detected as follows:

(a) Fundamental - 1260 Mc. Strongly evident through 50-db attenuation.

(b) Second Harmonic - 2520 Mc. Blanked by a high-power S-band radar operating near this frequency; thus, it was difficult to ascertain whether the second harmonic was actually present.

(c) Third Harmonic - 3780 Mc. Readily evident.

(d) Fourth Harmonic - 5040 Mc. Not detected.

(e) Fifth Harmonic - 6300 Mc. Readily evident.

(f) Sixth Harmonic - 7560 Mc. Not detected.

(g) Eighth Harmonic - 10,080 Mc. Not detected.

NOTE 1: The GRR-9 antennas were rigidly mounted to one plate so that individual antenna polarization rotation could not be achieved. All antennas were operating vertically polarized. The GRR-9 receiver was swept at 15 and 60 cycles with no apparent change in signal reception.

NOTE 2: The sensitivity to pulse signals were seen to increase by (1) increasing the intensity, and (2), by adjusting the vertical control so that the base line was not visible.

NOTE 3: It was determined that the difference between one scale division of local-oscillator noise (as set by the IF gain control) and three divisions of local-oscillator noise changes the system sensitivity by only 2 db. A calibrated signal generator was used to conduct this measurement at 8875 Mc. Thus, it is unnecessary to adjust the IF gain control to produce noise output greater than one scale division when looking at most signals.

d. Test Conclusions - The AN/GRR-9 receiver is effective in intercepting and observing the transmitter harmonics which are of a greater level than receiver tangential sensitivity. No interchannel feedthrough is obtained.

2. Test No. 2 - Signal-Intercept Frequency and Amplitude Measurements

a. Test Objective - To determine the effectiveness of, and the problems resulting from the use of the "signal-substitution method" to measure frequency and signal level on a swept receiver.

b. **Test Procedure**

   (1) The van site was initially located approximately 10 miles from Verona on Route #31. At this site, signal reception from the AN/FPS-8 was not achieved. The van was moved to a new site approximately 6 miles from Verona, on Route #31, on the old highway. Signal reception from the FPS-8 was achieved at $275°$, on the FPS-8 azimuth indicator. Signals were received so strongly at approximately 1300 Mc that the FPS-8 operators were requested to reduce the power output.

   (2) The AN/FPS-8 was the only operational radar and it was directed at the van ($275°$).

   (3) The vertically polarized GRR-9 antennas were directed at the radar.

c. **Test Results**

   (1) Test No: 2

   (2) Date: 10/18/62

   (3) Recorded By: W. Brown, AEL

   (4) Observers: A. Sparacino ⎫
                  A. Jackson ⎬ RADC
                  D. Synder   ⎪
                  F. Acquino ⎭

                  K. Walker, AEL

   (a) The intercepted signal was observed on the display; film Counter Nos. 1424-1527.

   (b) The output of an HP614A signal generator was inserted in the signal path through a 10-db directional coupler.

   (c) As shown in film Counter Nos. 1528-1564, the HP614A signal generator frequency was adjusted to give a zero beat with the AN/FPS-8 signal.

   (d) The amplitude of the HP614A signal was made equal to that of the AN/FPS-8 signal. As shown in film Counter Nos. 1565-1608, the signal-level indication, including line and coupler losses, was -67.8 dbm. No change was observed between the 15- and 60-cycle sweep.

   (e) The AN/FPS-8 signal was measured with the Polarad FIM Field Intensity Meter.

   (f) The frequency of the AN/FPS-8 was measured as 1302 Mc.

   (g) The power indication on the FIM-B2 read 42 db above 1 microvolt.

   NOTE: The calibration of the FIM-B2 meter indicated a signal strength of -74 dbm for a signal 42 db above 1

microvolt. However, one microvolt is equal to -107 dbm, and 42 db above this level is -65 dbm. Therefore, it was assumed that the FIM power indicator was out of calibration.

(h) The HP614A signal generator was set to an output frequency of 1314 Mc. This signal was fed to the FIM-B2. When the FIM-B2 meter was peaked, a frequency of 1302 Mc was indicated.

d. Test Conclusions

(1) A variable sweep width would have been desirable during initial interception of the AN/FPS-8 radar signal. Since the approximate frequency was known (and not the approximate azimuth direction) the sweep could have been reduced to a small portion of the frequency band and could have been centered about the AN/FPS-8 frequency. This procedure would have greatly increased intercept probability.

(2) The signal-substitution method for calibration of frequency and power on the GRR/9 receiver is therefore accurate in frequency to within 1 Mc and accurate in power to within 2.8 db of the FIM meter. The signal-substitution method is very easy to use for frequency measurement. Power measurement is somewhat more difficult since the received pulse amplitude changes slightly from pulse to pulse, depending upon the position of the pulse in the IF passband. In any case, the signal-substitution method is as easy to use as the FIM meter for both frequency and power measurements.

3. Test No. 3 - Field Surveillance Operation (Single Radar Intercept Aspects)

a. Test Objective - To investigate the effects of radar pulse repetition rates (PRF) and radar-antenna rotation rates (RPM) on the intercept capabilities of a swept receiver as functions of receiver display sweep rates and the screen persistency of the cathode-ray tube.

b. Test Procedure - Test No. 3 was run immediately after the conclusion of test No. 2, utilizing the AN/FPS-8 and the same physical configuration. Those present at Test No. 2 were present at Test No. 3.

The AN/FPS-8 radar is only capable of a 365 cycle PRF. However, rotation rates could be varied between 0 and 10 RPM.

c. Test Results

(1) 1 RPM - A signal was received from the radar approximately each revolution. Sometimes this signal was displayed as signal plus image, sometimes as signal alone, and other times as image alone.

(2) 3 RPM - Signals were received at a rate which was somewhere
between one intercept every other scan and two intercepts
out of three scans.

(3) 7 RPM - Sometimes signals were received for a period of
approximately 30 seconds, and at other times, two or three
intercepts in a row were achieved at approximately 9-second
intervals.

(4) 10 RPM - Four or five antenna cycles sometimes failed to
produce one signal intercept, as shown in the film, Counter
No's. 1740-1810. Counter No's. 1608 - 1739 are blank frames
showing base line presentation only. No appreciable differ-
ence in intercept frequency was noticed between the 15- and
60- cycle sweep rates.

A long-presistance scope was substituted for intercepts
in the L band. Although the information was retained for
a longer period of time when separate pulses were received
together, short, single pulses did not write on the scope.
However, with the lights turned out so that the van was
completely dark, short-pulse indications could also be seen
and the long-persistence scope increased the sensitivity
to some degree. Variable-PRF tests will be run at a later
date at AEL, and variable-PRF signal generators will be used
for these tests.

d. Test Conclusions - A paper-tape recorder or other visual recorder
would be valuable in the interception of radiation from scanning
antenna. When pulse interceptions are far apart in time, the
eye of the observer tires and a blink or sideways glance may
cause a received pulse to be missed.

4. Test No. 4 - Field Surveillance Operation (Multiple Radar Intercept)

a. Test Objective - To observe the operation of the AN/GRR-9 type
swept receiver in intercepting and indicating the presence of
multiple radar signals.

b. Test Procedure - Test No. 4 was run immediately after the con-
clusion of Test No. 1, and the same physical configuration was
utilized. Those present at Test No. 1 were present at Test No.
4.

c. Test Results - L-, S-, and C-band radars were operated simul-
taneously during this test. Although no control over the radars
was available, every signal on the indicators was able to be
accounted for, including a seventh order harmonic of the L-band
radar shown on the X-band indicator.

All radar signal levels were strong, since the van was
located close to the radar. The AN/GRR-9 antennas were vertically
polarized. All signals were clear and no interband feedthrough
was present.

d. <u>Test Conclusions</u> - The AN/GRR-9 receiver showed capability in intercepting and indicating the presence of multiple radar signals. As several of the signal levels were very high, excellent resistance to interchannel feedthrough was evident.